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Design and Realisation of Microwave Circuits

1. GENERAL STATE OF THE ART

If we look at the radical changes of the last twenty years in relation to electronics and the digital transmission and processing of analogue signals, it has often looked as if the end had come for the "classic communications and high-frequency technology".

But, upon closer analysis it becomes clear that what we are actually dealing with here is a structural change, for it is only the classic gauging equipment, transmitters and receivers with rotary knobs, pointer gauges, linear and circular scales and cylinder dials which have actually disappeared, along with the thick cables and the high-frequency plugs. The wave-guides have also disappeared in the frequency range above 1 GHz (apart from certain areas in which their use can not be dispensed with).

1.1. Review of State-of-the-Art in the Microwave Area

a. Entertainment electronics ("TV satellite dishes") have driven the development of components with less and less noise forward in the GHz range (e.g. GaAsFET's with noise factors below 1dB at 12 GHz) at ever more favourable prices and with millions of components being produced. Parallel to this development, more and more refined, and also more and more advantageously priced, circuit, design and tuning techniques came into being for GHz converters and receivers.

b. The further development of mobile radio telephones into ever higher frequency ranges and their simultaneous widely-developed marketing as a consumer product has brought more and more powerful transmission, mixing, modulation, oscillation, synthesiser and receiver IC's onto the market at ever-lower prices. Hand in hand with this, there has been further

development of the long-familiar transmission technology as a time-division or frequency-division multiplex method, combined with new types of modulation (usually phase shift-keyed) and signal processing techniques (e.g. DSP).

c. "Normal telephone communications" and data services (ISDN, etc.) are continually increasing their transmission speed and / or channel loading, and are suddenly having to use "electronic circuits" again (e.g. within the optical waveguide systems which are used now) which, at elementary frequencies which can be above 1 GHz, have to be built in accordance with the rules of high-frequency technology.

1.2. Microwave hardware

a. The system resistance (= impedance level, Z, of coax cable, etc.) is 50Ω all over the world.

b. In consumer electronics (TV satellite technology), the "F plug" dominates, together with double-screened coax cable with best damping values of - for example - 0.4dB per metre at 2 GHz, but with a system resistance of 75Ω.

c. In commercial technology in the frequency range going up to 18 GHz, the SMA plug represents the predominate standard for the connecting up of assemblies which do not have to be continuously assembled and dismantled; for this there are, for example, SMB plug connectors. The N-plug has certainly not been paid off completely, but is being pushed further and further into the background.

d. The "high-frequency connection lines" of assemblies normally consist of semi-rigid cables. These are small, semi-rigid copper tubes, with Teflon filling and a solid, thin, silver-plated or gold-plated internal conductor. In the SMA plug, the internal conductor of the cable is used directly - pointed and then used as the central pin of the plug and socket connection. These cables are bent into the desired form during manufacture, and should not need to be altered again.

   The two types "RG 405" and "RG 402" are normally used here. They have different external diameters, with the damping on the "RG 402" cable (external diameter 3.58mm) being about 1.95dB per metre for the frequency f = 18 GHz, while by contrast that for the "RG 405" cable (2.7mm external diameter) is 3.75dB (at 1 GHz: 0.36 or 0.67dB per m).

e. For flexible connections - depending on the frequency range and the damping requirements - coaxial cables with varying degrees of insulation and diameters are used. In the application range above 1 - 2 GHz, we find Teflon (PTFE) used exclusively as a dielectric, to keep the losses low.

f. The circuits themselves are realised almost exclusively by means of the SMD technique, with fibreglass-reinforced epoxy resin printed circuit boards being used up to app. 1.5 GHz (FR4 material). As the

frequency rises, the transfer is made, either to fibreglass-reinforced Teflon (RT-Duriod, Ultralam 2000, DiClad, etc.) or ceramic substrates (Al_2O_3) or to artificial sapphire.

g. CAD software, which has had a long time to mature, with astonishing efficiency and precision, exists for the circuit design work (including all the calculations for antennae and filters). The interested amateur, instructor or engineer naturally profits by this, since this software is often also on the market in shareware versions at more favourable prices. Here we shall mention only the Puff CAD software, which was actually developed for training purposes in a high-frequency institute.

2. PROPERTIES AND APPLICATIONS OF PUFF

The Puff CAD program can be used to analyse normal circuit structures or strip transmission line circuits and then to optimise them.

But things go fastest if the PC has a mathematical coprocessor to increase the power, which increases the speed approximately fivefold.

Puff determines the scatter parameters (= S parameters) for a pre-selected frequency range, and plots the calculation results, either in complex form in the Smith diagram or in the well-known frequency response form (= rectangular plot).

The reference resistance of the system can naturally be pre-set. 50Ω is standard, but 75Ω is normal in Europe for low frequencies and TV satellite installations.

If desired, from the frequency domain response, Puff can determine the impulse response or, optionally, the time domain response, with the help of the fast Fourier transformation, and plot it in a diagram of its own.

The individual calculation results are stored in a table and can be displayed for every frequency step, in accordance with value and phase, by moving the cursor along the curve.

The Smith charts or frequency responses displayed can be printed out using Printscreen.

To optimise the circuit, an individual circuit parameter (e.g. the value of a component or the length of a strip line, etc.) can be infinitely varied within any limits, which you can select yourself.

Example for altered AUTOEXEC.BAT file is show overleaf.

Finally, the circuit layout can be printed out, to various scales (standard case: somewhere between m = 1:1 and a magnification with m = 5:1).

```
@LOADHIGH C:\DOS\SHARE.EXE /L:500 /F:5100
C:\DOS\SMARTDRV.EXE
@ECHO OFF
PROMPT $P$G
PATH C:\GEOWORKS;E:\MSIMEV54;E:\WINDOWS;D:\WINDOWS;C:\DOS;C:\SKETCH3
C:\DOS\NLSFUNC
SET TEMP=C:\DOS
C:\DOS\MODE CON CODEPAGE PREPARE=((850 437) C:\DOS\EGA.CPI)
C:\DOS\MODE CON CODEPAGE SELECT 850
C:\DOS\KEYB GR,,C:\DOS\KEYBOARD.SYS
C:\DOS\GMOUSE.COM
E:\WINDOWS\WIN.COM
```

Modified AUTOEXEC.BAT File

3. INSTALLATION OF PUFF ON A PC

First Step:

A directory called Puff is set up on the hard disc, and all the files from the diskettes are copied into it.

Second Step:

Before starting this program, you must switch over to the American character set ("code page 437") (otherwise there is no sign, etc.). To do this, it is necessary to alter the AUTOEXEC.BAT and CONFIG.SYS files. So the code set change must be carried out before Puff is loaded, using the DOS command "chcp" (= change code page).

Example for altered CONFIG.SYS file is shown below.

Third Step:

If the layout of the printed circuit board or the complete screen with all the data and diagrams has to be printed out, the

```
LASTDRIVE=Z
DEVICEHIGH=C:\DOS\SETVER.EXE
DEVICEHIGH=C:\DOS\HIMEM.SYS
DEVICEHIGH=C:\DOS\EMM386.EXE RAM
DOS=HIGH
COUNTRY=049,850,C:\DOS\COUNTRY.SYS
DEVICEHIGH=C:\DOS\DISPLAY.SYS CON=(EGA,,2)
FILES=40
STACKS=9,256
SHELL=C:\DOS\COMMAND.COM C:\DOS\ /P
BUFFERS=40
```

Modified CONFIG.SYS File

appropriate resident printer driver must be loaded (just as before the start of PUFF.EXE). In each case, 2 drivers are supplied for: HP Laserjet, Epson and IBM Proprinter (for EGA or VGA screens).

They are called:

ega2eps.com vga2pro.com

vga2eps.com ega2lasr.com

ega2pro.com vga2lasr.com

If you want to make life simpler, then write the following small batch file - "PUFF.BAT" - and store it in the master directory of the C drive (which will be selected when the computer is switched on). Then you only have to call PUFF-.BAT up, and everything else takes care of itself.

```
chcp 437
cd c:\puff
vga2pro.com
puff
chcp 850
```

4. S-PARAMETER

A brief introduction to the S or scattering parameter

It is not so easy to measure flows and voltages at high frequencies, moreover open-circuit and short-circuit measurements can no longer be made correctly (e.g. to determine the internal resistance of a source). The system description and calculation should therefore be based on values which can also be easily measured.

For this reason, another working hypothesis is basically already in use today for frequencies from 50 to 100 MHz:

The central point is a system resistance which is the same everywhere (here, for example, 50Ω), which applies to the input and output resistances, the impedance level of the cable, and the moving loads.

The measurement technique now concentrates on, for example, determining the resistance values present (or their deviations from 50Ω) - in complex form, of course, i.e. according to amount and phase.

a. We begin at the signal source with the system resistance as internal resistance - here, 50Ω.

b. Our subject - for example, a filter, a moving load, an amplifier, a mixer or an antenna - is connected to this source through a relatively long cable with Z = 50Ω.

c. With longer cable lengths, due to the final propagation speed of 30cm per nanosecond in air, nothing is noticed immediately from the consumer.

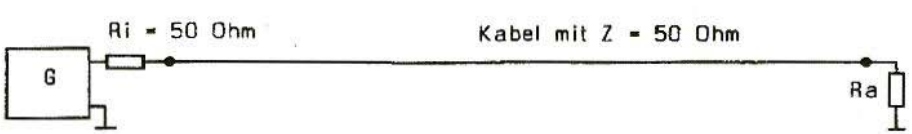

Fig. 1: 50Ω System; *Kabel mit* = Cable with

So the cable displays an input resistance of 50Ω, and forms a voltage divider with the internal resistance of the source. Thus, at first we have "power matching, $R_i = R_a$" and the active power picked up goes on its way in the direction of the consumer at "cable speed":

$$V_{cable} = C / \sqrt{\varepsilon_r}$$

d. If this incoming power (correctly described as an incoming wave) reaches the consumer, then it is fully "absorbed" only if power matching occurs here too, i.e. $R_a = Z = 50\Omega$. If there are deviations from the power matching, the "surplus" is returned to the source, and then a "return wave" can be observed on the line from the load resistor in the direction of the source.

e. For this reason, the concept of the reflection factor, "r" is introduced, which gives us:

$$r = U_{rück} / U_{hin}$$

or return voltage:

$$U_{rück} = r \times U_{hin}$$

where

$$r = (Z_{last} - Z) / (Z_{last} + Z)$$

A voltage at the load is then simply:

$$U_{last} = U_{hin} + U_{rück}$$

Note: for these moving waves, Ohm's law must naturally apply all over the line.

$$U_{hin} / U_{hin} = Z (= 50\Omega)$$

and

$$U_{rück} / U_{rück} = Z (= 50\Omega)$$

More about the S-Parameters:

Energy transmission on the line naturally always occurs through outputs. But in order to be able to work with the transmitted outputs using expressions in which the voltages appear, we simply take the square root out of the output formulae and christen the result the "wave value".

This gives the incoming wave value, a:

$$P_{hin} = (U_{hin}^2/Z) > a = \sqrt{P_{hin}} = (U_{hin}/\sqrt{Z})$$

and correspondingly, for the return wave value, b:

$$P_{rück} = U_{rück} > b = P_{rück} = (U_{rück} / \sqrt{Z})$$

The significance of these actions does not become clear until we consider a quadripole - e.g. an amplifier! For this quadripole will, for example, have an amplification value, but there are also return effects from the output on the input.

First step:

If the input is controlled by a signal source, this is described as the incoming wave value, a_1. If we now consider the "Echo b_1" arising in the feed line to the input with a directional coupler, then it consists of 2 parts, namely:-

- the reflected fraction of a_1, which is created by the deviation of the input resistance from $Z = 50\Omega$, and
- a second fraction, which is generated at the input by feedback from the output, where indeed some signals, a_2, are present

So for the echo:

$$b_1 = (a_1 \times S_{11}) + (a_2 \times S_{12})$$

Significance of coefficients:

If the output is precisely matched, we can not expect echoes to be reflected back from there, and the value, a_2 becomes zero. We thus obtain S_{11}:

$$S_{11} = b_1 / a_1 \qquad \text{for } a_2 = 0$$

This is none other than the input reflection factor, r, from the previous calculation example! However, in practise it is always complicated, due to running times and reactive components. Thus, for example, value and phase are always given in relation to the frequency in transistor tables.

The value S_{12}, then, is the "reverse transfer ratio". It supplies data on the feedback values for the quadripole if the input is not controlled and is also matched with Z = 50Ω. In this case, the output is "blown on" by a signal generator with the output a_2:

$$S_{12} = b_1 / a_2 \qquad \text{for } a_1 = 0$$

Second step:

Now we apply the same train of thought to the output. We imagine a signal source on the right, which processes the output of the quadripole with the wave value a_2. We then obtain, as an echo, b_2:

$$b_2 = (a_1 \times S_{11}) + (a_2 \times S_{22})$$

If the input of the amplifier is just not controlled, a_1 is automatically zero. For this reason, a "signal coming from the quadripole output" can arise only through a reflection at the internal resistance of the quadripole.

Thus, S_{22} is the output reflection factor of the quadripole!

$$S_{22} = b_2 / a_2 \qquad \text{for } a_1 = 0$$

Now only the left-hand summand in the equation remains, and this is already known. If a signal is transmitted at the input alone and the output is simply matched, with Z = 50Ω, we obtain the classic amplifier circuit. S_{21} is now nothing more than the "voltage amplification of the quadripole with correct matching at the output".

$$S_{21} = b_2 / a_1 \qquad \text{for } a_2 = 0$$

Practical tips:

The resistance values can be determined from the input and output reflection factors in the following way. (Please take note that we are dealing with complicated values here!)

For example input resistance:

$$Z_{11} = (Z \times (1 + S_{11})) / (1-S_{11})$$

or internal resistance:

$$Z_{22} = (Z \times (1 + S_{22})) / (1 - S_{22})$$

If we have the resistance values and wish to use them to calculate the reflection factors (see our example above), then, using the well-known method, we obtain:

At the input:

$$S_{11} = (Z_{11} - Z) / (Z_{11} + Z)$$

At the output:

$$S_{22} = (Z_{22} - Z) / (Z_{22} + Z)$$

You can save yourself these conversions by working with the Smith diagram!

To be continued

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Design and Realisation of Microwave Circuits Part-2

5. THE AUXILIARY EQUIPMENT

The aim here is to give a brief sketch of the equipment needed for the "microwave workshop", so that even a beginner can have the opportunity of finding out about the most important things. Not only do microwave circuits behave differently to short-wave circuits - or VHF circuits - they usually also need a different mechanical structure. Precision mechanics is often a pre-requisite here for good electrical measurement results.

5.1. The Auxiliary Mechanical Equipment

We have already referred in the introduction to the need to use SMD technology for the entire assembly. Fig.1 shows the most important auxiliary equipment:

Fig.1: Appropriate Aids make the Work Easier

- Temperature-controlled soldering irons with very narrow replaceable soldering tips
- Extra-fine watchmaker's tweezers
- Stereo magnifying glass in form of spectacles
- Set of SMD chip resistors from 10 Ohms to 1 MOhm, component size for beginners: 1206 (3mm x 1.5mm); component size for those more experienced: 0805 or 0604 (2mm x 1.25mm or 1.5mm x 1mm)
- Set of SMD chip capacitors from 1pf to app. 100nf - the capacitors from 1pf to 220pf should be size 0805, suitable for microwaves (e.g. "ultra-stable NPO" from Philips / Valvo, or in the Siemens "COG - Class 1")
- Roll of SMD solder (100g is enough!) with 0.5mm wire diameter
- Pivoted precision vice with ball and socket joint or printed circuit board holder.

Upright drilling machines, vices, files and taps, or perhaps a small lathe, will be ideal for the mechanical operations.

5.2. Electronic Auxiliary Equipment

The electronic auxiliary equipment includes, for example, digital multimeters, and at least one adjustable DC voltage power pack with adjustable current limiting. Something new is the requirement for a PC - if possible from a 386SX upwards, with a coprocessor and Windows. (Why Windows? All will be explained...)

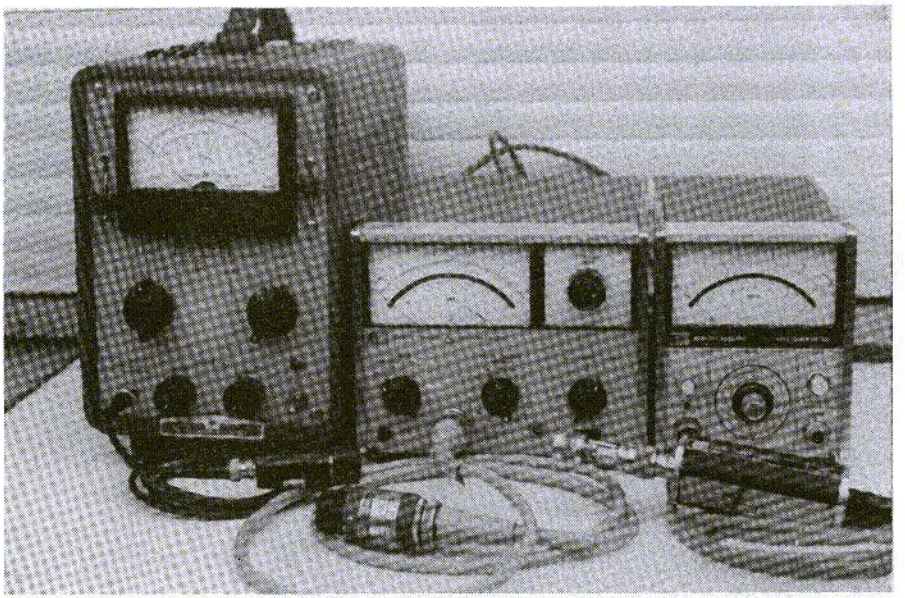

Fig.2: Examples of Relatively Old High-Frequency Measuring Equipment

The following are useful aids for the microwave measuring equipment:

a. Test transmitter for the intended measurement range, from 50 Ohms upwards, with a 50 Ohm internal resistance. It's quite OK if the test transmitters date from the valve era, but they should be equipped with a good reducer covering a wide range of levels - e.g. 120dB.

b. People like to use the wobbler range from Hewlett-Packard, which sell in large numbers. They are easy to maintain in working order, and it is also easy to obtain documentation about them. These units can also be used as test transmitters in the "CW" position.

c. At least one microwave power meter- but better two, with measuring head for up to, for example, 12 GHz. They can be used to obtain very precise measurements of input and output power levels, and thus of the amplification or attenuation brought about by a module (i.e. our S-parameter S21).

   Fig.2 shows examples of the most widely used equipment - from left to right.

- The HP430A (to 430C) is undoubtedly the "grandmother" of entire generations of wattmeters. Although developed back in the fifties, it is precise enough. One unpleasant fact is that this "bolometer meter" can be used to measure only very low power levels of about -20dBm (20mV can be measured at 50 Ohms).

- The next generation of meters, 431 and 432, are already greatly improved. The measuring heads have temperature compensation and the zero point can be corrected.

- The well-known Powermeter 435A has a chopper amplifier in the measuring head for the minute DC voltage output of the detector. Using a thermo-element instead of a thermistor, we can now measure as far as -65dBm.

d. It is scarcely possible to operate without high-frequency plug connectors, adapters and high-frequency cables of various lengths. Here one should stock up with appropriate materials covered by the "N" and "SMA" standards, in which context preference should be given to used or corroded MIL quality instead of high-gloss nickel-plated Taiwanese products. Naturally, equipment may sometimes have to be bent back into shape during cleaning, to restore a good earth contact again.

   SMA connectors are very good for operating up to 18 GHz, but nevertheless attention should be paid to the following points:

- In practice, an SMA connection is made only once, and here the union nut is tightened at 56Ncm, using a special dynamometric key. Thus the manufacturers do not see it as a connection which can frequently be undone.

- For this reason, if they are being used in experimental rigs (with a lot of screwing processes), they should

be treated with great care. Check them regularly for damage, and clean them even more regularly.

e. You can't do without a small set of coaxial moving loads, short-circuits and attenuators, with values between 3dB and 30dB. "N" type formats are good but bulky. In contrast, "SMA" formats are incredibly small. Special attention should be paid to good quality here, as otherwise you'll be puzzling over the reasons for poor matching values in the measuring rig later.

f. For those also wishing to measure reflection factors, there is the not exactly cheap directional coupler, with 20dB decoupling attenuation and a directivity exceeding 30dB. Luckily these articles can usually be picked up as surplus material, with N or SMA plug connectors.

g. Just one more thing, learned from personal experience. You work hard and long and very successfully with the equipment referred to above. But inevitably there comes a day when you decide that you can't get any further without a spectrum analyser with a frequency range, for example, up to 18 GHz. Only the price scares you off! However, there are two possibilities here - either buy a used SA, or create a DIY apparatus, following the assembly instructions of - for example - J.Jirmann, DB1NV.

6. METEOSAT CONVERTER APPLICATION PROJECT

A specific project makes it possible:

- To plan a system and lay down specifications
- To draft circuits and analyse them, with the aid of appropriate microwave software (e.g. Puff)
- To develop the final current cycle plan and convert it into printed circuit boards, using CAD software if necessary (e.g. Geddy-CAD shareware)
- To design and manufacture appropriate housings, suitable for microwave work, for the assemblies
- To dimension the individual modules and the apparatus as a whole, and to document the measurement readings

Using the new design aids naturally offers other options as well. The modules created step by step in this way can also be put together extremely rapidly in a rather different way - one thinks of a 1,296 MHz assembly, or an assembly for the GPS receiver. The additional development cost would then extend over only a few stages, which would have to be "re-dimensioned". In most cases, this is done by calling up the appropriate Puff file and entering new design data.

It is also possible to put together a complete circuit from individual part-circuits already developed, which can then be fitted onto an overall printed circuit board.

6.1. Framework Conditions

Fig.3 shows the overall circuit diagram for the Meteosat converter. The only striking feature here is that the two MMIC amplifiers (described below) are identical.

Fig.4 shows the specimen rig for the complete converter. In the test layout, the converter was housed in the vicinity of an offset mirror with a DIY exciter. It was remotely powered through a coaxial cable, about 10m long, connected to the output.

The following pre-conditions apply to the module design:

a. The individual stages are housed in milled aluminium housings with screw-on tops, have 50 Ohm input or output resistors, and are fitted with SMA jacks on the narrow sides.

b. The smallest printed circuit board size is 30mm x 50mm, with a housing measuring 35mm x 55mm x 15mm high. If this can not be achieved, the board size can go up to 30 x 95mm and the housing size to 35 x 100mm. The fixing system on the board consists of 4 M2.5 x 8 screws, each screw being 3mm from the edge of the board (this gives a "bore pattern" of 24 x 44mm).

c. The operating voltage is fed into the interior of the housing through a feedthrough capacitor, to an SMD voltage stabiliser (type 78L05).

d. The internal core of the SMA jacks must be fed through the housing wall so as to "match" the impedance level, and must then be connected to the board using a 50 Ohm microstrip line, with as little reflection as possible.

Fig.5 shows a blow-up view of these details on the pre-amplifier module.

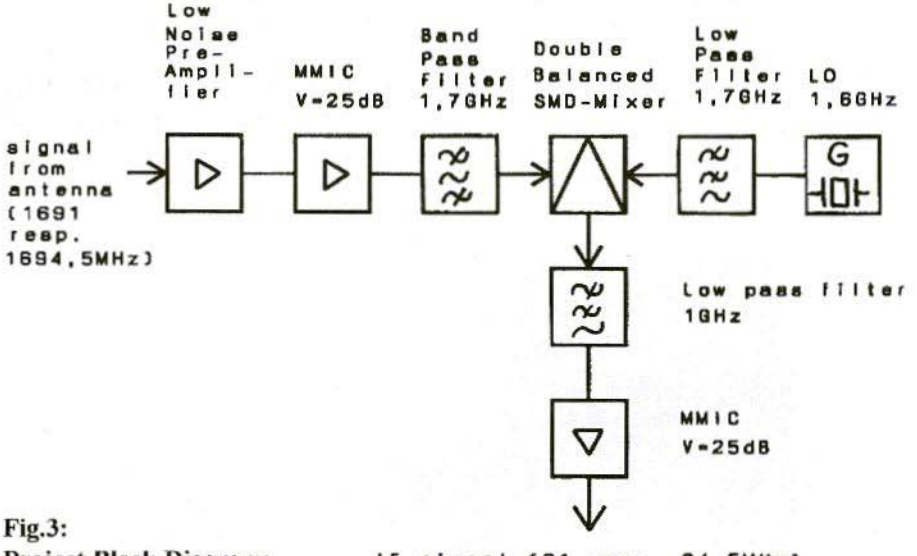

Fig.3:
Project Block Diagram:
Meteosat Converter

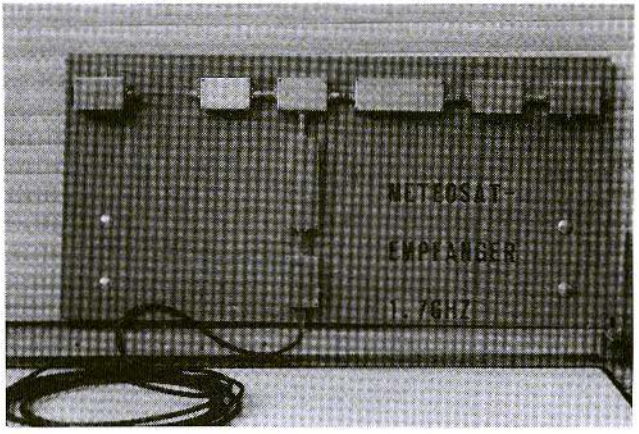

Fig.4:
Specimen Converter System Assembly

6.2. The first Assembly: LNA for 1.7 GHz

6.2.1. The Basic Circuit of the LNA

To obtain a low overall noise factor, the noise factor of the input amplifier must be kept as low as possible, and the amplification factor must be as high as possible, so that the subsequent stages can not have a stronger effect. The state of the art here is the GaAsFET, and the choice fell on the CFY30, for reasons of price. We can learn from its data sheet, or from the semi-conductor data CD-ROM from Siemens, that a minimum noise factor of 1 to 1.2dB is possible at 1.7 GHz. Associated with this, of course (see Fig.6, left-hand diagram), is a "source resistance" at 1.7 GHz of Zi = 100 Ohms + j150 Ohms.

The right-hand diagram of Fig.6 shows, for the 4 GHz frequency, how accurately this transformation must be carried out to keep the noise low. Thus, if you imagine the central point of the concentric circles at our point, 100 Ohms, being displaced by only + j150 Ohms, you obtain the circumstances at 1.7 GHz.

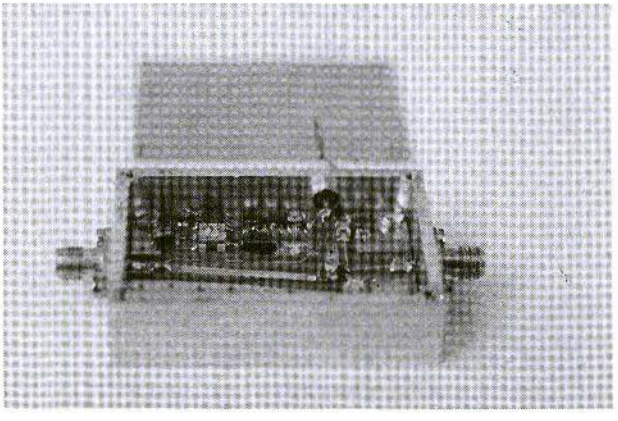

Fig.5:
Details of Pre-Amplifier Assembly

At this point, we should mention something else which can have a significant influence on design. If the exciter does not give an internal resistance of precisely 50 Ohms at 1.7 GHz, then the transformation circuit is displaced in relation to our start point. Things become dramatic if the aerials and the amplifier input are connected by coaxial cable! In that event, the start point actually moves - depending on the cable length - around a circle centred on the centre of the diagram. It is easy to imagine how far you can suddenly go from the Fmin point. The craziest thing about all this is that the noise factor will suddenly periodically oscillate between 1 and 10dB if you alter the cable length.

Once the aerial resistance has been established at - hopefully - exactly 50 Ohms, we have to try to make this into 100 Ohms + j150 Ohms at the gate of the FET. This can be done in several ways. One thinks of stubs, transforming L/C networks, strip lines wired in series, etc.. After tests on the various possibilities using Puff, and in view of their influence on the frequency cycle, however, the following solution was selected.

The 50 Ohm resistance of the aerial is initially transformed to 100 Ohms with a lambda/4 strip line. For this purpose, the strip line must have a wave resistance of:

$$Z_{strip} = \sqrt{(50\Omega \cdot 100\Omega)} = 70.7\Omega$$

At 1,700 MHz, the missing j150 Ohms call for a coil with 15nH. There are lovely, tiny little SMD inductances for this, size 0805 (i.e. 2mm x 1.25mm). At this frequency, the coil quality is given as "min. 50", so it is completely adequate.

We can naturally not leave the gate connection in limbo, so we feed the required DC, in a known fashion, into a further highOhm $\lambda/4$ line, short-circuited at the end (with a wave resistance of at least 100 Ohms). Due to the transformation into an open circuit, the FET, hopefully, does not notice what we have done.

For higher amplification, the output side of the FET is subjected to communications adaptation. To do this, the drain DC is fed in through an SMD inductance, the reactive impedance of which is high, as against the internal resistance of the FET. 47nH was selected, giving j470 Ohms. With an appropriate transformation circuit - e.g. a strip line, once again - you must also ensure that the reflection factor is zeroed on the SMA output jack. Puff can be of assistance here.

The basic circuit from Fig.7 is thus slowly crystallising, and we can risk our first simulation using Puff.

6.2.2. S-Parameter Files

To simulate a given circuit, we need the S-parameters of the semi-conductors used. In our example, this means the specifications for the recommended operating point from the data sheet (e.g. drain voltage 3.5V, drain current 15mA) and a file with the S-parameters for this operating point. We can come at this in two ways:

a. We can obtain the appropriate diskettes with S-parameters for all standard semi-conductors - here then,

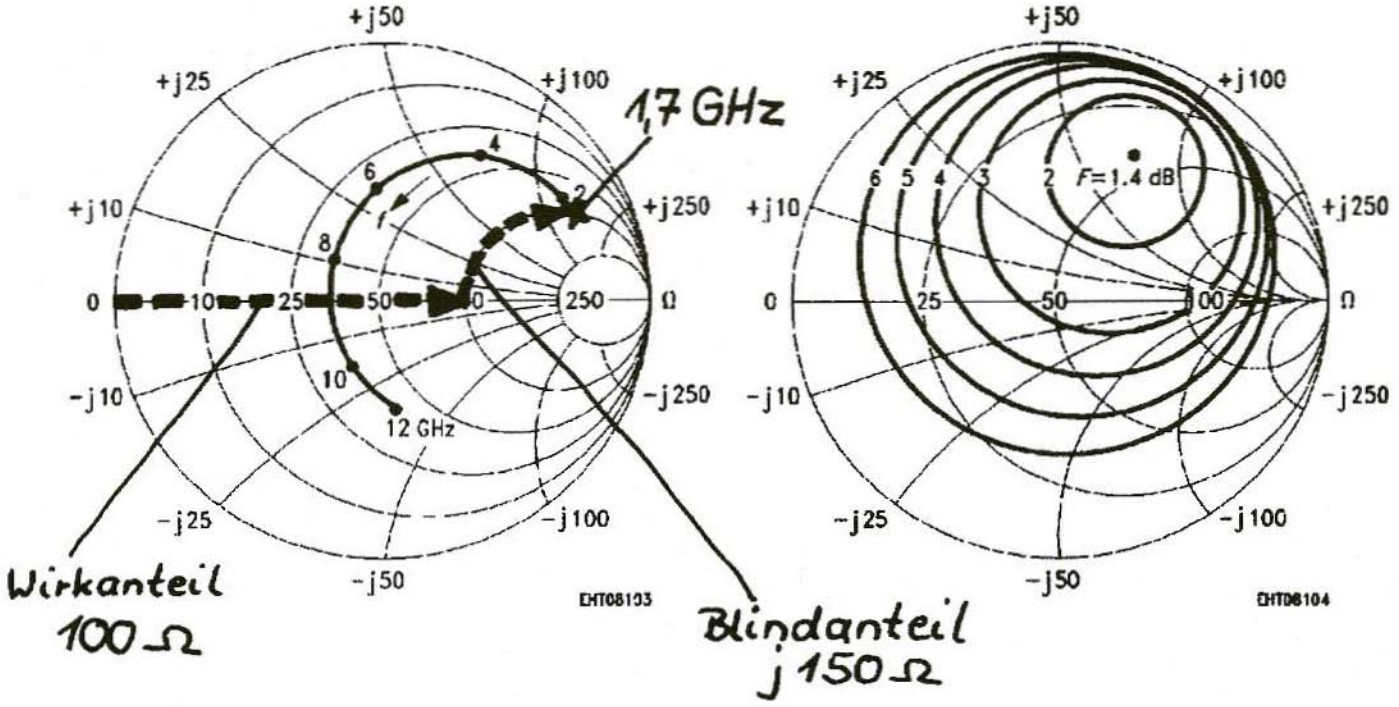

Fig.6: Diagrams
Wirkantiel = Active Fractions; Blindantiel = Reactive

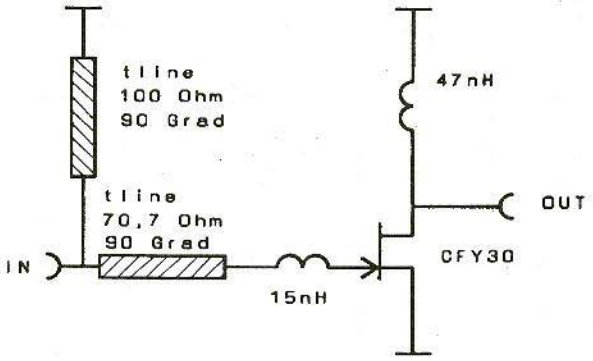

Fig.7:
Base Circuit of LNA

the Siemens high-frequency transistors. The data for the semiconductor in question - here the CFY30 file - still have to be adjusted, as the data are usually supplied in "touchstone format" (identification: *S2P), which Puff can not read. There is a conversion program for Siemens data, which is also supplied on the Puff diskette.

b. There are also data on CD-ROM - e.g. from Siemens. This CD-ROM also contains a table with S parameters, in the appropriate data sheet. These data must also be adjusted. The simplest way is using a word processor. Any kind of "device file" supplied with Puff with the ending "*dev" is loaded into the word processor, and this is used as a pattern for how such a file must be structured. Thus, you simply modify all the specifications for the CFY30, the S values always being: first amount, then phase angle. Finally the file is saved into the "Puff" directory, here with the name "CFY30.dev".

N.B.: Most touchstone files do not have any values for the "zero Hertz" frequency. There is no alternative to estimating them, on the basis of the other values, and entering them into the tables subsequently! Puff otherwise sends an error message and goes on strike when the simulation should begin at zero Hertz.

6.2.3. The Printed Circuit Board material

Now all we need for the first simulation are the data on the board material used. On grounds of price, and since we are still operating below 2 GHz, the "FR4" fibreglass epoxy material was selected - photo-coated on both sides with a thickness of 1.5mm and a 35mm copper deposit.

Of course, there were considerable difficulties at first:

- No values relating to the dielectric constants for this material could be obtained from the supplier. In the specialist literature, we found the value "5.5", so a small aerial intended for 1.7 GHz was designed for the test. To our amazement, the resonance frequency turned out to be

f	S11	S21	S12	S22
0	1 0	2.23 180	0.01 90	0.69 0
1	0.98 - 18	2.23 157	0.024 74	0.69 - 15
2	0.95 - 39	2.22 136	0.045 57	0.66 - 30
3	0.89 - 64	2.21 110	0.068 40	0.61 - 45
4	0.82 - 89	2.23 86	0.085 23	0.56 - 62
5	0.74 - 115	2.19 61	0.099 7	0.49 - 80
6	0.65 - 142	2.11 36	0.107 - 10	0.41 - 98
7	0.58 - 171	2.05 11	0.113 - 26	0.32 - 119
8	0.51 155	1.99 - 15	0.117 - 41	0.23 - 147
9	0.50 120	1.85 - 40	0.115 - 56	0.16 167
10	0.51 90	1.62 - 65	0.110 - 69	0.17 118
11	0.53 65	1.44 - 86	0.107 - 80	0.25 84
12	0.55 41	1.34 - 107	0.107 - 92	0.31 60
13	0.59 17	1.20 - 130	0.104 - 105	0.38 35
14	0.64 - 3	1.03 - 153	0.099 - 118	0.45 16
15	0.68 - 19	0.86 - 170	0.092 - 131	0.51 - 1

Table: Here is an S-Parameter File as an Example
CFY30 GaAsFET in SOT143 package
V DS = 3.5V, I D = 15mA, Z 0 = 50 Ohm
frequency f in GHz

almost 2 GHz, so this value could never be correct. Moreover, the static capacity of the large printed circuit board supplied was measured with a capacity measuring bridge, and the dielectric constant was determined from this as being 4.8.

- The next trial using the 1.7-GHz aerial still produced unacceptable discrepancies, so once again we looked for the reasons. There follows a summary of the results:

- FR 4 material has a dielectric constant of 4.8 at low frequencies;

- This value decreases, from about 1 GHz upwards, first slowly, but then faster and faster; for 1.7 GHz, it is only 4.3.

- The losses likewise rise slowly from 1 GHz, but then faster and faster. At 1.7 GHz, we can expect a value of 0.015 for the loss factor.

- Unfortunately, the dielectric constant varies by up to 3 per cent from delivery to delivery. In critical applications, such as resonators or aerials, for every new board material delivery, the dielectric constant variation must first be determined, and must be taken into account in the Puff layout, on a percentage basis, when the value is pre-set.

- Anyone for whom this is too expensive should immediately change to a decent microwave printed circuit board material with a Teflon or ceramic basis, such as RT-Duroid, etc.. These materials conserve their characteristics practically unaltered

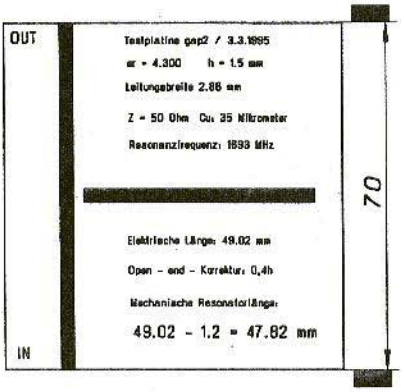

Fig.8:
Aerial for 1.7 GHz as Test Rig

Testplatine **= Test board;** ***Leitungsbreite*** **= Line width;** ***Mikrometer*** **= Micrometres;** ***Elektrische Länge*** **= Electrical length;** ***Open-end-Korrektur*** **= Open end correction;** ***Mechanische Resonatorlänge*** **= Mechanical resonator length**

up to and above 10 GHz, and the manufacturer's specifications can be transferred unchecked into the Puff simulation.

If, on the contrary, you stay with FR4, then you can first engrave yourself a test circuit - for example, as per the specimen in Fig.8. In the test circuit, a stub cable at no-load is connected through a small 0805 chip capacitor to a continuous 50Ohm line, which is closed off at 50 Ohms. For all λ/4 resonances of the stub cable, this forms a short-circuit, and thus dramatically worsens the otherwise good input reflection factor. At the same time, the layout acts as a band barrier at this frequency. Using these two effects, it is then possible to work backwards over the mechanical length of the stub cable to determine the actual dielectric constants at this frequency.

6.2.4. Entering Puff

Now we come to Puff. The procedure is explained step by step.

1. Start Puff by calling up the "PUFF-.BAT" file.

   Hopefully, this will also put the "Code Page 437" into operation (command "chcp 437"), to change over to the American code set with the Ohm and degree signs. If not, the line

   ```
   DEVICE=C:\DOS\DISPLAY,SY
   CON=(EGA,,2)
   ```

   must be added in CONFIG.SYS and the lines

   ```
   C:\DOS\NLSFUNC
   MODE CON CODEPAGE
     PREPARE=((850 437)
   C:\DOS\EGA.CPI)
   MODE CON CODEPAGE
     SELECT=850
   ```

   must be added in UTOEXEC.BAT. Of course, the three lines must be set in AUTOEXEC.BAT as far as possible at the beginning, or else before the driver for the keyboard, CD-ROM, etc..

2. Press key F3 to arrive in the "Components Window".

3. The components required in the circuit are now entered one after another.

   – Discrete components always start with "l(umped)", followed by the component value

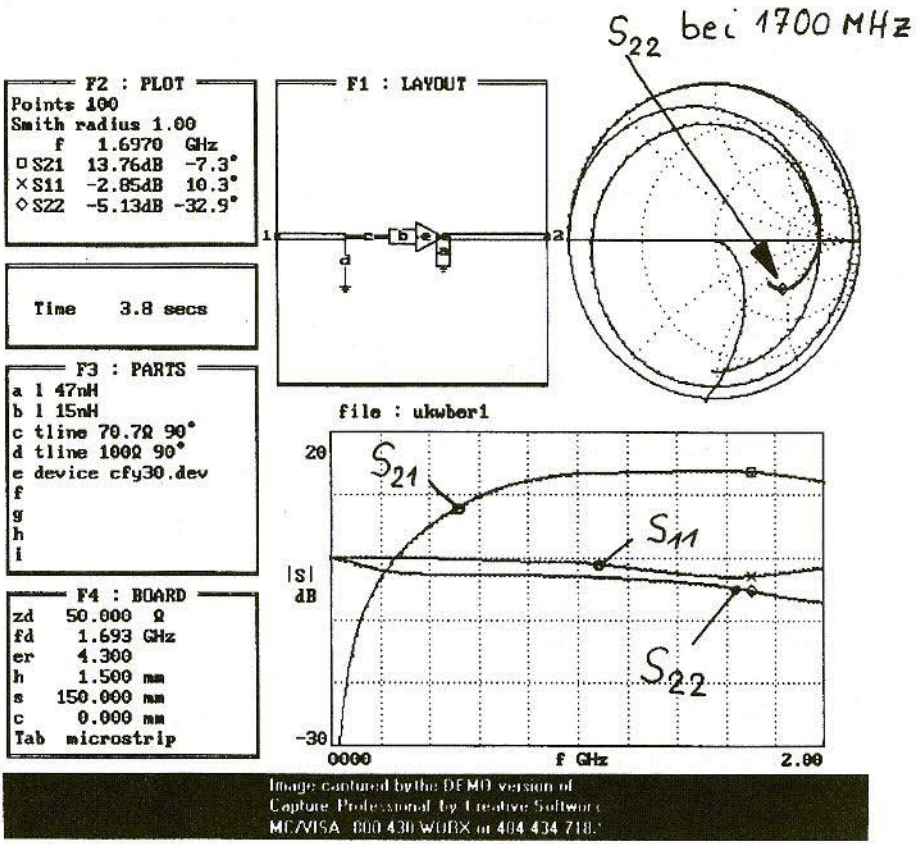

Fig.9: Puff Screen Print-Out; *Bei* = At

- Strip lines begin with "t(line)", followed by the impedance level and the electrical length in degrees
- Active components are entered using "device *.dev"

Fig.9 shows how the complete list for the specimen circuit should look.

4. Next, press key F4 and enter the values for impedance level (50 Ohms), mean frequency (1.693 GHz), dielectric constant (4.3) and board thickness (1.5mm). 150mm should be the board size selected.

5. The key F1 takes you to the layout window, i.e. to the board. If an old design can still be seen there, it can be deleted using "CTRL e". The "start cross" is then displaced by pressing the change-over key and moving the appropriate cursor key slightly to the left, to obtain space for the circuit.

   If you now press key "c", this component (it is the 70.7Ohm transformation line) is active, and can be placed, using the appropriate cursor key.

This continues to the end of the procedure, during which:

- A wrongly positioned or superfluous competent can simply be deleted again by pressing the change-over key and "Cursor Key in Opposite Direction"
- You can travel round the circuit already "plotted" with the cursor at any time without any consequences, and
- Upon completion, press "1" to connect up to the input jack
- Press "2" to connect up to the output jack

  But take care; within the circuit, the cursor must also actually be at the input or output point of the circuit when the corresponding key is pressed, otherwise you will connect up some point or other of the circuit. In the worst case, you can delete the circuit using "CTRL e" and begin again.

6. Activate key F2 to branch into the calculation program. First you preset, for example, 100 calculation points, and select only sizes S21 and Sss. At the other S parameters, you need only delete the index, so that they are "switched off". Use the "Up" or "Down" cursor key to get back to the axes of the diagram in which the results are displayed. On the frequency axis, you key in, e.g., "0" and "2.0 GHz", and on the vertical axis the values "-30dB" and "+ 20dB" are shown.

If everything is now in order, press "p" to obtain the screen as in Fig.9. If problems arise, compare the values for all the windows of the print-out with those on your own screen. You'll soon find the errors. You can see the beautifully even amplification curve given by the cycle of S21, going from 1 GHz almost all the way to 2 GHz.

Now, of course, we still need to adjust the power at the output. For this, you should cast a glance at the Smith diagram of the print-out and have a look at the cycle of S22. The curve runs to the right below the mean point. This means that the internal resistance of the FET exceeds 50 Ohms and has a capacitive blind fraction.

It can be transformed to 50 Ohms only by means of a strip line, which has:

A resistance considerably exceeding 50 Ohms, and:

Is considerably shorter than 90 degrees, so that you are in the vicinity of the 50Ohm point in the Smith diagram.

Thus, as an experiment, you might enter a strip line having a value of Z = 120 Ohms and a length of 30 degrees, which you position between the FET output and the output terminal "2" into field F3. Moreover, the 47nH SMD choke coil must still be connected somewhere on the drain connection of the CFY30 subsequently, at size 1206 a gigantic component in comparison to the FET. So we also need a small piece of strip line here, with as high an impedance level as possible (that gives greater inductance) and a length of app. 5mm. But things more or less have to stop at Z = 135 Ohms, since then the width of the line is already falling below 0.3mm.

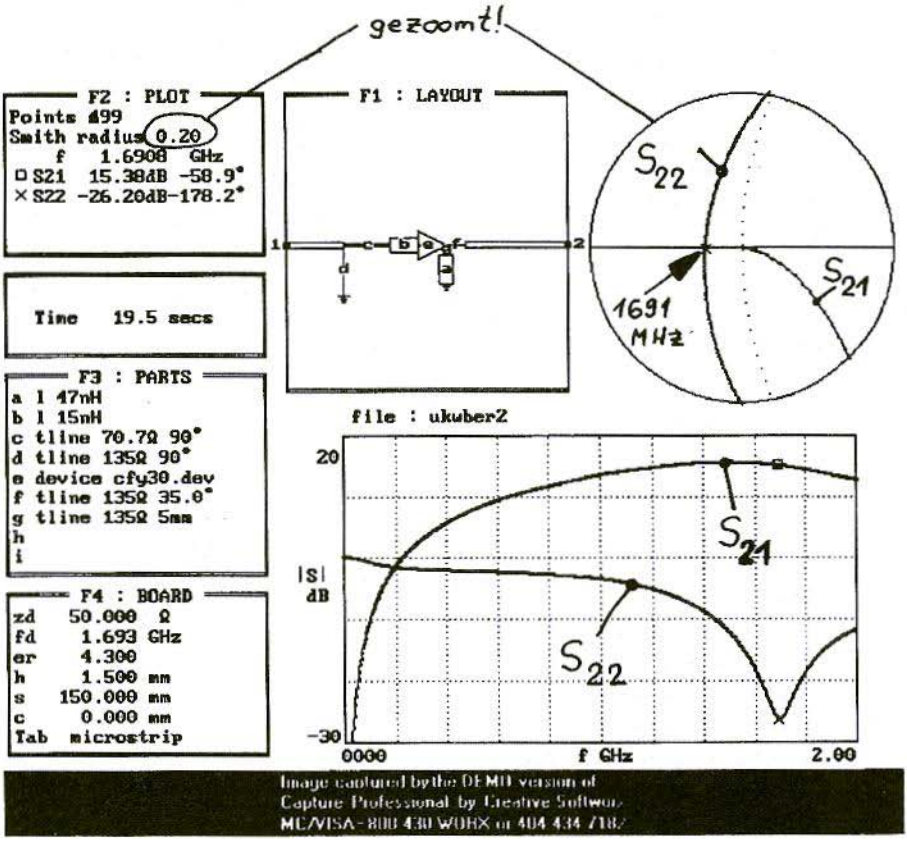

Fig.10: Optimised Circuit Data; *Gezoomt* = Zoomed

(Even for the gate pre-voltage for the operating point, we therefore select this 135 Ohms as impedance level for the lambda/4)

We now go back to field F1 and incorporate the changes just made into the circuit. Next, we need only to make a new plot, to examine the results in the diagrams, and to correct the line length in the direction of point 2. The entire process will need to be repeated a few times, until you can get practically as far as possible from the mean point (= 50 Ohms) at 1,700 MHz in the Smith diagram and are simultaneously on the vertical axis. S22 will then display a minimum value! If the S22 curve finally runs past the 50 Ohm point to the left, then the transformation resistance must be slightly increased again. If, on the contrary, it runs past to the right, then the resistance must be slightly decreased.

One other recommendation here - you should "zoom out" the mean point of the Smith diagram, so that you can fiddle around until the S22 curve is as near to this point as possible, and thus the best

possible matching is precisely at 1,693 MHz. For this, we need not only to reduce the "Smith radius" in field F2 right down to 0.5, or even 0.2, but we should also increase the number of plot points (absolute limit: 499). Hopefully, your PC has a coprocessor. Otherwise, with circuits of considerable size, the calculations will take a very long time.

Fig.10 shows the data for a circuit optimised in such a way.

Before continuing, we must not forget to carry out an essential Puff check. Change the frequency range scanned into "0 to 10 GHz" and carry out another plot. You now need razor-sharp eyes to investigate the cycle of S11 and S22 in the bottom frequency diagram, as the values of S11 and S22 must never exceed 0dB anywhere.

Those would mean reflection factor values exceeding 1, and these are unfortunately associated with negative resistance values for the input resistance and also for the output resistance. But in most cases anything like that causes the amplifier to oscillate.

Now one more observation which I have recently made concerning the print screen function:

It has become clear that, of the printer drives supplied on the Puff diskette, only the "Epson version" functions without any problems on all printer versions at all times. Printing with HP Deskjets and Laserjets has proved to be very much the biggest headache in the past. Part of the screen was always missing, and there was no way to eliminate the errors. Even a letter to the software authors in America resulted only in an answering letter containing the priceless final sentence:"...let us know if you find a solution, because this is a common problem...".

The lengthy searching and experimentation has borne fruit, for I can now offer an almost perfect solution. By chance, in a sub-category on a CD-ROM belonging to my children (111 Windows Games), I found a "Screen Capture Shareware Program" named "Capture". This was the solution to the problem, and printing out became child's play.

Install Puff using Windows (as a DOS application by means of the program manager) and start it up from there. If you are working with Puff and you press the "Printscreen" key as usual, the screen is copied into the intermediate filing (= Clipboard). Go back to Windows, call up "Capture" and load the intermediate file into this program. The many processing options now on offer make it a pleasure to invert (to eliminate the black background), convert into a black and white image, balance the half-tones, rotate the diagram, and much more.

You can save the end result as a "Bitmap file" and print out immediately, using the PC system's Windows printer. Naturally, I prefer to fetch the image into a "graphics frame" in my word processing program (Winword 6.0) and then cut the bottom 2cm off, as that's where the tiresome reference to the shareware version appears.

You should now become familiar with the first Puff operating commands. Unfortunately the original English manual does not contain a list of all possible commands and operations. Since one is

really helpless without this reference work, I have compiled the following "Alphabetical Keyword Catalogue for Operating with Puff".

6.2.5. Alphabetical Keyword Catalogue for Operating with Puff

Command Keyword	Key Combination
Indicate dimensions of components or lines:	Go into field F3, move cursor to corresponding component line and press "="; data then appear in information field
Load American code set (to access Ohm sign, etc.)	Before calling up PUFF.EXE: "cd c:\puff", then "chcp 437" (N.B.: First change AUTOEXEC.BAT and CONFIG.SYS!)
Place component (in layout F1)	Change to field F1, move cursor to intended connection point and select component by keying correct letter in (see list for field F3!). Then place component by pressing corresponding cursor direction key.
Delete component (in layout F1)	Change to field F1, move cursor to start of component and delete it with key combination "Shift + cursor direction key"
Enter component values (in F3)	Change to field F3, move cursor tolist empty space in list and key in: "l(umped)". The components can be entered individually, as a series circuit, or as a parallel circuit. For Ohmic resistors, either "Ohms, kiloOhms, etc." or values standardised on the system resistance (1z", "0.3z", etc.) are permissible as units for Ohmic resistors. Blind elements, on the other hand, can be given directly with the correct unit (e.g. pf, mH), or as imaginary resistance values ("+j10"), or as imaginary resistance values standardised on the system resistance ("-j5z"). Example for series circuit: "2+j4-j3z".

Expand component list F3, because points a...h are not sufficient	Press F3, then activate tab key
Component sweep	A component from the component list can be varied in its value through the program, whilst the other components and the frequency are kept constant. In this way, a circuit can be optimised. To do this, first go to field F3 and insert a question mark in front of the component value. Example: lumped ?10pF or tline 50 ?90° After the change to field F2, the sweep limits for the variation of the component are set on the axes of the frequency cycle diagram - for example, 5pF to 15pF for the capacitor or 70° to 110° for the stripline.
Insert coupled lines	Enter in component field F3 with "clines" or just "cl", followed by values of "even" and "odd" resistance, if impedance level of coupled lines is different from system resistance. Then enter electrical length. If impedance levels of system and coupled line are the same, only one of the two resistance values is to be entered (either "Even" or "Odd" can be used). Puff then calculates the absent second value and enters it in the *puf file.
Load file of a draft	Change to field F3, "CTRL r", enter file, then "Return".(N.B.: File must have extension *puf. But this extension does not have to be indicated.)
Save file of a draft	Change to field F2, then "CTRL s", key in file name, "Return".
Change board material	Go into field F4, then enter new thickness value under menu point "h"

Print out draft layout	Change to field F2, then "CTRL a" (Enlargement factor p must have already been set again, if necessary, through word processing, directly in circuit file *puf!)
Load printer driver	Before calling up PUFF.EXE: "cd c:\puff", then e.g. "vga2pro.com" for IBM Proprinter; then "puff.exe"
Degree sign (°) for entering angles for components, e.g. for electrical lengths of line units	In field F3 "Alt d"
Request help	Press key "F10"
Trigger pulse answer	In field F2 "i"
Micro-sign (e.g. for micro--Henry)	In field F3 "Alt m"
Use micro-stripline with dispersion and attenuation	Insert tline! instead of tline. N.B.: Before Puff is loaded, the file in question (or the set-up file, for a new start) must be called up with a text editor, and the board thickness, the thickness of the Cu layer, the surface roughness and the loss factor for the intended draft frequency must be entered and saved. Example: values for FR 4 material for the frequency 1,700 MHz er = 4.3 loss tangent = 1.5 E-2 Board thickness s = 1.5mm Conductor thickness = 0.035mm Surface roughness = 0.001mm
Jump to next node in layout	In field F1 "CTRL n"
OHM sign	In field F3 "Alt o"
Delete one component on board	Change to field F1, move cursor to start of component and delete it using key combination "Shift + cursor direction key"
Delete complete circuit on board	Change to field F1, then "CTRL e"

Parallel sign for discrete component combinations	In field F3 "Alt p"
Board: Change interval of	Change to field F4, then key in new portpins value under menu point "c"
Board: Change external dimensions	Change to field F4, then key in new value under menu point "s"
Board: Change thickness	Change to field F4, then key in new value under menu point "h"
Board: Change dielectric constant of material	Change to field F4, then key in new value under menu point "er"
Board: Create earth connection at a circuit point	In field F1 "="
Change number of plot points	Change to field F2, move cursor to "Points" and enter number of calculation steps (max. 449) manually
Start plot procedure (including calculation)	In field F2 "p"
Start plot procedure after change and also show old cycle	In field F2 "CTRL p" a
Connect portpin to layout	Change to field F1; move cursor to corresponding end point of circuit, key in number of portpin, then "Return"
Load Puff	Suggestion: Batch file "PUFF.BAT" into basic directory c: otherwise: "chcp 437", "cd c:\puff", "vga2pro.com", "puff"
Leave Puff	Press "ESC" twice in succession
Smith diagram: Change cut-out radius	Change to field F2, move cursor to "Smith radius", then key in desired radius. (Values between zero and infinity are permissible. Values smaller than 1 give a "cut-out enlargement".)
Smith diagram: Switch over to "Screen filling"	Change to field F2, then enter "Alt s"
Smith diagram: Print screen	Press "Printscreen" key (N.B.: This is not effective unless a printer driver - e.g. VGA2PRO.COM has been loaded before the start.)

Smith diagram: Switch back from "Screen filling" to "Small representation", with separately represented frequency cycle curve:	In field F2 "ALT s"
Smith diagram: Displace frequency markers	See under "Frequency markers". Displacement with "Page Up" or "Page Down"
Smith diagram: Switching over from resistance representation to electric conductance representation	Go into field F2, then activate "Tab" key
Trigger jump answer	In field F2 "s"
Change start frequency for plot procedure and calculation	Change to field F2 and move cursor key ("Up" or "Down") until cursor is on start frequency of lower diagram (for frequency cycle). Key in new start frequency, then start new plot procedure with "p".
Change stop frequency for plot procedure and calculation	Change to field F2 and move cursor key (up or down) until cursor is on stop frequency of lower diagram (for frequency cycle). Key new stop frequency in GHz in, then start new plot procedure with "p".
Select S parameter for calculation and plot procedure	Change to field F2, go down in field F2 with cursor key and key in desired parameter after letter "S" (e.g. S11)

6.2.6. Expansion of Basic Circuit to full Wiring Diagram

To obtain a circuit which can be assembled, the following requirements must also be met:

- The drain source DC voltage must be + 3.5V, the drain current must be 15mA
- The gate must have a negative voltage, app. 0.35V for this current, against the source connection
- The source connection itself must be at earth potential for high frequencies
- The mains supply voltage for the FET amplifier should be +5V, and is obtained from "Remote power supply at +12V" by means of an SMD fixed voltage regulator (type 7805 / SMD)

There are now 2 options for setting the operating point via the Gate voltage:

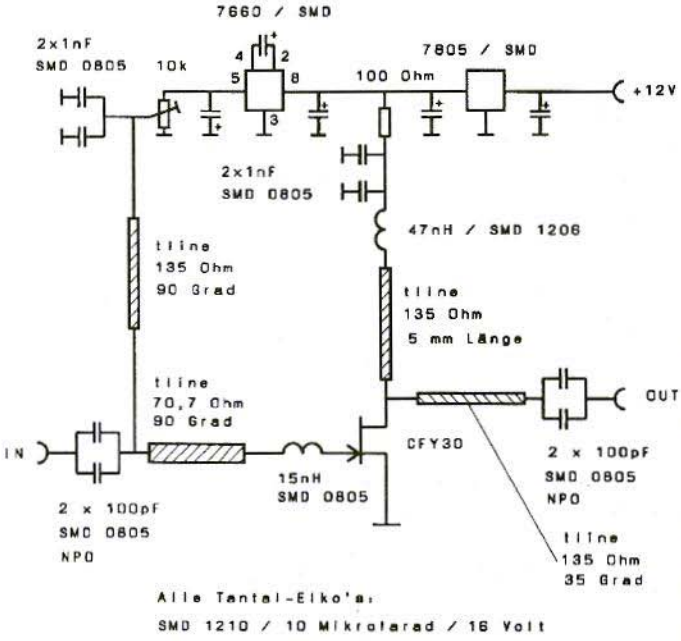

Fig.11:
Wiring Diagram with Negative Bias Voltage Generation

Alle Tantal-Elko's **= All tantalum electrolytic capacitors**
Grad **= Degree**
Länge **= Length**

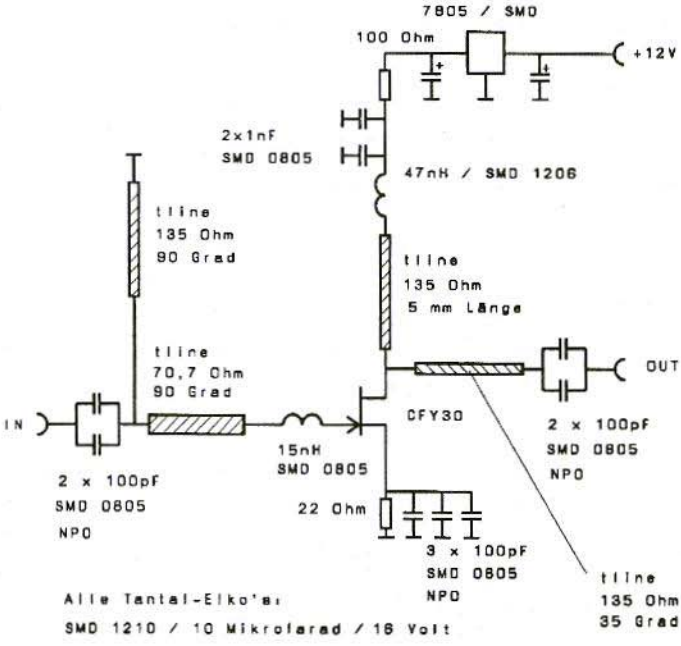

Fig.12:
Circuit with Source Resistance Solution

1. An SMD switching controller (7660) is used to generate a negative voltage of -5V, which is used to power both the known lambda / quarter line and the gate, through a read-out potentiometer. The source can now be directly connected to the earth.

2. The gate connection is put directly at earth potential through the lambda / quarter line. But here a 22 Ohms resistance is needed in the source feed, on which the desired negative gate pre-voltage is set. This is the simpler solution, but it causes considerable difficulties for broad-band, lowOhmic earthing of the source connection at up to 2 GHz using appropriate capacitors.

To investigate these influences in detail, I simply set both options up and then took measurements.

Fig.11 shows the assembled circuit with negative pre-voltage. Fig.12, on the other hand, shows the solution using the source resistance.

(To be continued)

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Design and Realisation of Microwave Circuits

Part-3

6.2.7. Board Layout for both Versions of the Circuit

For a project on a board, we also need the width and the length of the striplines used. But for this purpose we also need to carry out a few more simulations:

Puff in fact also supplies us with the behaviour of the circuit for "non-ideal striplines", in which the losses through the board and the skin effect are taken into account, together with the dispersion (= scattering, i.e., for example, the emergence of lines of electric flux at the edges of the stripline and the consequent impedance level changes, etc.).

Here, experience shows that the more accurate the material data are the more precisely the simulation will reflect reality, and the discrepancies can be reduced to less than 1%!

For this purpose, the Puff file for the amplifier circuit to be processed has first to be loaded into the word processor and the board material data have to be entered. For the FR4 material with a thickness of 1.5mm. and 2 x 35µm.

\b{oard}	{.puf file for PUFF, version 2.0}
d 0	{display: 0 VGA or PUFF chooses, 1 EGA, 2 CGA, 3 One-color}
o 1	{artwork output format: 0 dot-matrix, 1 LaserJet, 2 HPGL file}
t 0	{type: 0 for microstrip, 1 for stripline, 2 for Manhattan}
zd 50.000 Ohms	{normalizing impedance. 0<zd}
fd 1.693 GHz	{design frequency. 0<fd}
er 4.300	{dielectric constant. er>0}
h 1.500 mm	{dielectric thickness. h>0}
s 150.000 mm	{circuit-board side length. s>0}
c 0.100 mm	{connector separation. c>=0}
r 0.010 mm	{circuit resolution, r>0, use Um for micrometers}
a 0.000 mm	{artwork width correction.}
mt 0.035 mm	{metal thickness, use Um for micrometers.}
sr 2.000 Um	{metal surface roughness, use Um for micrometers.}
lt 1.5E-0002	{dielectric loss tangent.}
cd 5.8E+0007	{conductivity of metal in mhos/meter.}
p 5.000	{photographic reduction ratio. p<=203.2mm/s}
m 0.600	{mitering fraction. 0<=m<1}
\	

Table 1

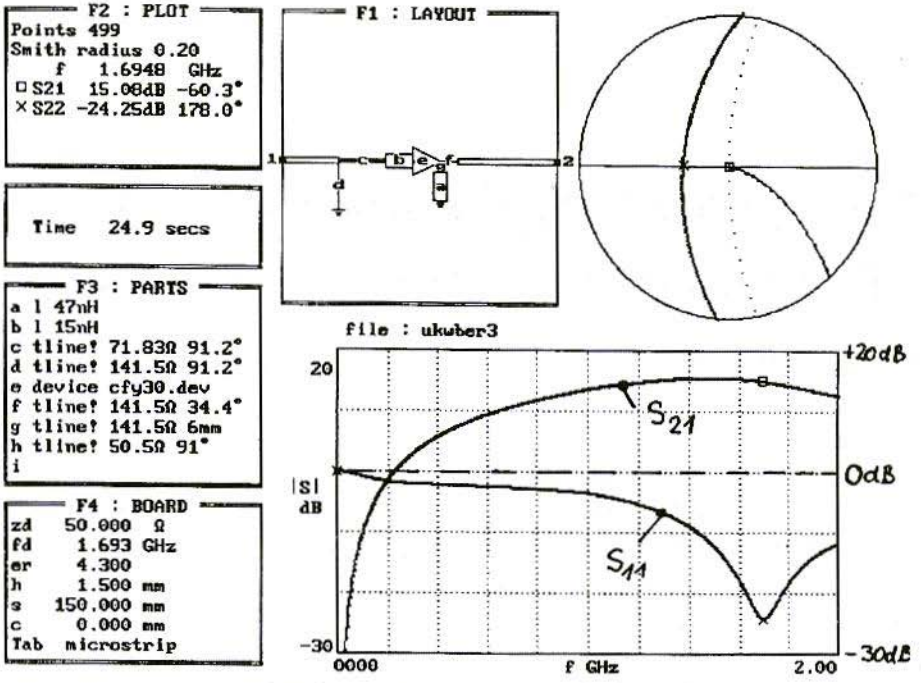

Fig.13: "Puff" Screen Print-Out: Ideal Amplifier with CFY 30

copper plating, it should look like this for 1.693 GHz - the important entries are marked out in bold and the rest of the file, with the S-parameters, etc. is left out (Table 1).

When the file has been stored, Puff is called up again and the amplifier circuit to be processed is loaded. In field F3, for each stripline, an exclamation mark, "!", is now placed directly after the entry "tl" or "tline". This switches to "advanced modeling", the most accurate reflection of reality.

But now there is more work to do. If the cursor is moved to each of the individual striplines in turn and the equals sign, "=", is keyed in, then the actual current characteristics of the line being scanned are displayed in dialogue field F1. It should come as no surprise if there are divergences, sometimes considerable, from the (ideal) value entered. But keep correcting the impedance level and the electrical length of the line in field F3 until the values in F1 or the frequency responses correspond to the ideal when plotted once more (call-up into field F2 with letter p)!

If this has been tackled correctly (and don't forget, the new circuit in field F2 has to be stored under a suitable name, using the command "CTRL s"), then we obtain a Puff screen in accordance with Fig.13.

The mechanical dimensions of the striplines corrected in this way, which are in fact needed for the board project, are obtained as follows:

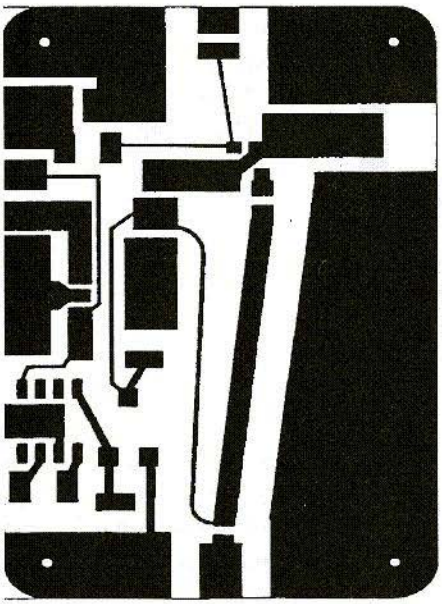

Fig.14: Printed Circuit Board Layout for Circuit with Directly Earthed Source Pin and Negative Gate Bias Voltage

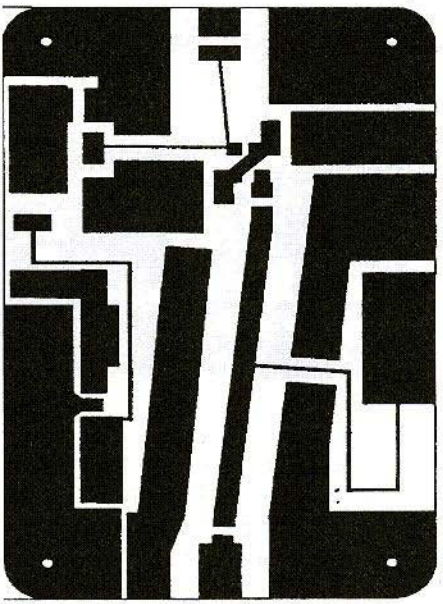

Fig.16: Layout of Circuit with Source Resistance and Capacitive Source Earthing

Move the cursor back to the line in question in field F3, delete the exclamation mark after "tl" or "tline" - for example, by pressing the space bar - and re-enter the equals sign. The length, l, and the width, w, of the selected microstrip appear immediately in dialogue field F1. Make a note of these readings and do not forget to re-enter the deleted exclamation mark in field F3 afterwards. Repeat the process for all lines to obtain the following list (Table 2).

Now we can move on to the actual project board. For this purpose, of course, we require a modern layout CAD system of some kind.

Component	Description	Z in Ω	El. Mechanical Length	Length mm	Width mm
c	Microstrip	71.83	91.20 Grad	25.301	1.504
d	"	141.50	91.20 Grad	26.597	0.221
f	"	141.50	91.20 Grad	10.032	0.221
g	"	141.50	20.04 Grad	6.000	0.221
h	"	50.50	91.00 Grad	24.614	2.874

Table 2

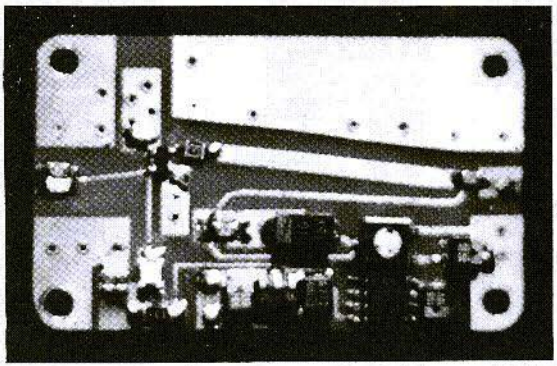

Fig.15:
Completed Prototype of Low-Noise Amplifier Direct Earthing

I am making no recommendations here, as there is a very wide choice. I myself still enjoy working with Geddy-CAD, which is frequently to be found on electronic CD's, on diskettes or in mailboxes. Once you have cleared the initial hurdles, it is extremely suitable for the direct drafting of Puff layouts. (Contact the publishers for more details on Geddy CAD.)

Fig.14 shows the finished layout. The board and the components for the circuit version with a directly earthed source pin and an IC 7660 for generating the negative gate bias voltage are depicted in Fig.15.

Fig.16 shows the layout of the circuit version with the gate earthed by means of a quarter-wave transmission line and a shorted-out source resistance. Fig.17 shows the prototypes.

To produce the boards, I use a somewhat unconventional but thoroughly practical method. The layouts are printed out onto transparent paper (greaseproof paper) by the laser printer on a 1:1 scale, and are used for direct exposure of the photo-coated FR4 material. Conductor widths going down to 0.2 mm. can be handled without problems. For higher precision requirements - for example, stripline filters or (small)

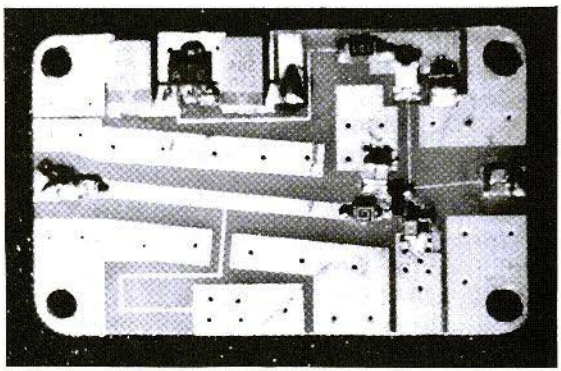

Fig.17:
Fully Fitted-Out Prototype of LNA Capacitively Earthed Source Pin

series production - the drafts are magnified before printing until they just fill a DIN A4 sheet. They are then printed out onto pure white paper with as smooth a surface as possible, and then this is filmed, with the film, again, showing the board on a 1:1 scale. In this way, printing errors during scaling down are greatly reduced, and we can get down as far as a 0.1 mm. track width.

The layouts themselves, of course, are the product of any amount of experience and likewise of many mistakes and discarded habits. So there follows a series of clarifications, which can also be consulted during the drafting of additional converter circuits:

a) Earth surfaces should be selected to be as large as possible. The minimum distance between them and the striplines should be 1 to 2 x the thickness of the board. Thus, firstly, the impedance level of the line is not too strongly influenced, and secondly the leakage fields of the lines can now no longer wander around so much in the region and cause the circuit to oscillate.

b) The earth surfaces of the input section (= gate area), the source area and the output section (= drain area) must be separated from one another, as otherwise there will be more and more of a tendency to oscillate due to couplings of one kind or another.

   These individual surfaces should now be connected to the continuous earth surface on the underside of the board in as low-impedance and low-inductive a manner as possible. In practice (e.g. with SAT-LNC's), this is done through many little holes which are then through-plated. I don't have this option, and found a solution in a Philips application circuit with silver-plated hollow rivets of 0.8 mm. diameter.

c) The source connection is particularly critical for the layout, since every effective or blind resistance has an immediate negative feedback effect here. The consequences can never be forecast precisely, but can be estimated using a Puff simulation:

   The amplification often falls more rapidly than expected in the upper frequency range, but just as often the input or output reflection exceeds 1 and the circuit oscillates!

   The manufacturer therefore also connects this electrode diagonally to two pins of the housing; firstly as a screen between the gate and drain, secondly to halve the impedance to earth, as now both pins are wired in parallel.

   Consequently, each source pin has its own earth surface, which must be earthed using at least 5 rivets, starting as close as possible to the pin!

   A Puff simulation can be used at any time to demonstrate that even changes of 1nH in the gain slope have a noticeable effect above 1 to 1.5 GHz!

d) When soldering on SMD components, you must ensure that these through-platings or hollow rivets do not become full of zinc, for any bore which is full is ineffective for high-frequency work, because of the low

penetration depth (= skin effect!) at these high frequencies the current path through this bore is blocked! In this way, unfortunately, the total inductance of the through-platings then increases again somewhat, and we give away the advantages we took so much trouble to obtain.

e) Should the circuit not fill the housing with its striplines, we also need a means of connection to the input and output plugs. This is brought about, naturally, through 50-Ohm striplines of appropriate length, so the widths of these lines are also in the list above.

f) Once capacitors become necessary in any high-frequency path, several SMD chip capacitors are connected in parallel, if possible. This again reduces the total inductance of the configuration, and the coupling or earthing becomes more broad-band.

g) Should it not be possible to gain control of the tendency of an amplifier circuit to oscillate, in spite of these measures, SAT-LNC's tried and trusted method using damping material made of conducting foamed material can be turned to on a trial basis.

These strips or panels of black foamed material are used as packaging material for CMOS-IC's. As a trial, a suitable piece is placed onto the circuit, and a test is carried out of how effective it is in damping. Subsequently, this material is glued to the inside of the housing cover (if gluing is necessary). Of course, there are very wide divergences in the effects of the various materials in circulation. For micro-wave engineering, the versions with low surface resistance are required, which can be recognised by the fact that they have a resistance of max. 10 to 100kΩ, measured at a distance of 10mm.. Moreover, these are usually the "harder" versions of the foamed material.

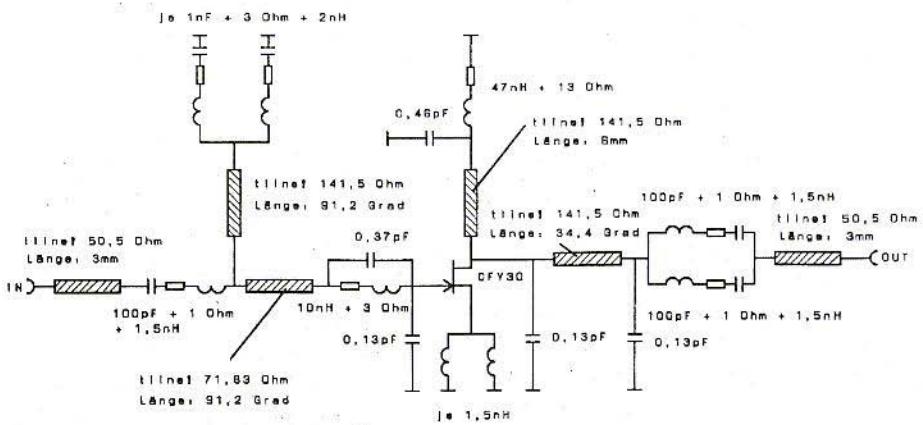

Fig.18: Advanced Simulation Circuit using Puff
***Länge* = Length**

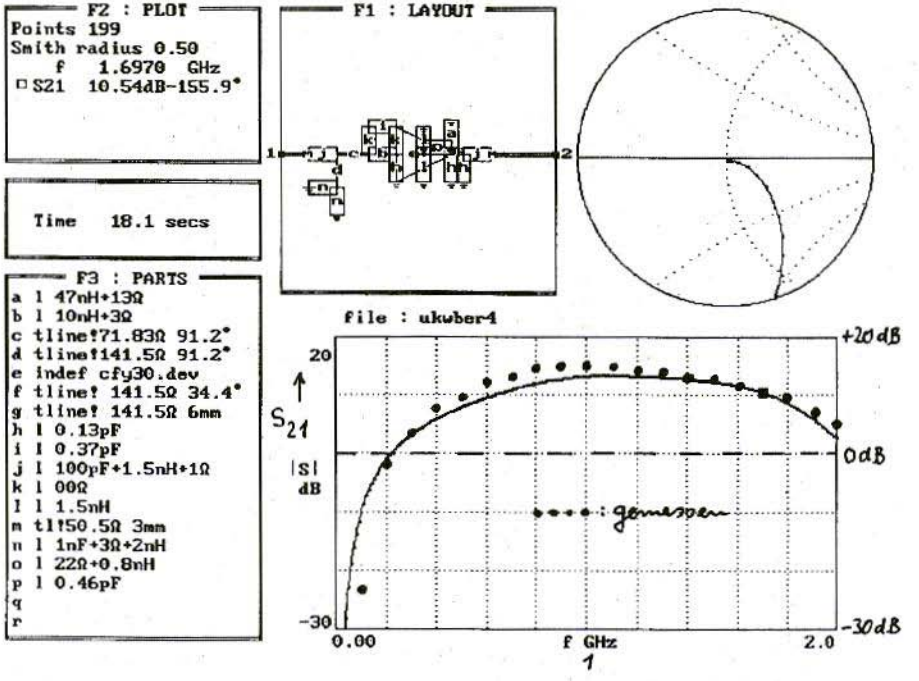

Fig.19: Puff Screen Print-Out: Simulation Result for Circuit with Directly Earthed Source Pin: Amplification I S21 I from 0 to 2 GHz

6.2.8. Simulation Results and Test Readings of the Version with Directly Earthed Source Pin

When we assemble the project circuit and measure its frequency response for the first time, we will certainly be greatly disappointed at the divergence from the simulation results. If we can gain control of the unintentional oscillation, that still does not mean the battle is won. The behaviour at high frequencies is not as desired, with the amplification falling relatively faster there. The "artificial model" must therefore be considerably expanded and refined to approach the reality.

The following facts, in particular, should be taken into consideration:

a) At all soldering points in the high-frequency stage which do not form part of a stripline, a capacity of app. 0.1pF must be provided for the "soldering scab" towards the underside of the board, i.e. the continuous earth surface. A value of 0.13pF produced the best agreement between theory and practice.

b) All chip earthing and coupling capacitors display a low inductance and also a series resistance, due to losses in the ceramic material. There is usually a diagram somewhere in the manufacturer's data books which gives the impedance gradient up to the maximum frequencies. The replacement values can be calculated

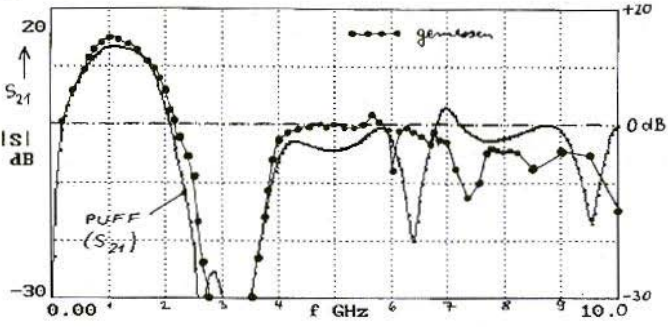

Fig.20: Comparison of Simulation Result and Measured Values for Circuit with Directly Earthed Source Pin in Frequency Range 0 to 10 GHz

from this, or from the self-resonant frequency and the rise in impedance above the self-resonance.

For the chip capacitors used at 10pF, made of the material NPO, a series circuit is thus obtained, as a sufficient approximation, from 10pF + 1Ω + 1.5nH.

c) The SMD coil, with 10nH at the gate of the FET, has a quality grade of 50 at 1,700 MHz. Due to the unavoidable self-capacitance, its inductance has already risen to 13 nH. According to the data sheet,its self-resonant frequency lies at "2,300 MHz". I have measured values of between 2,500 and 2,600 MHz.

Its replacement wiring diagram gives an inductance of 10nH in series with a resistance of 3Ω. A 0.37pF capacitor is wired in parallel with this series circuit.

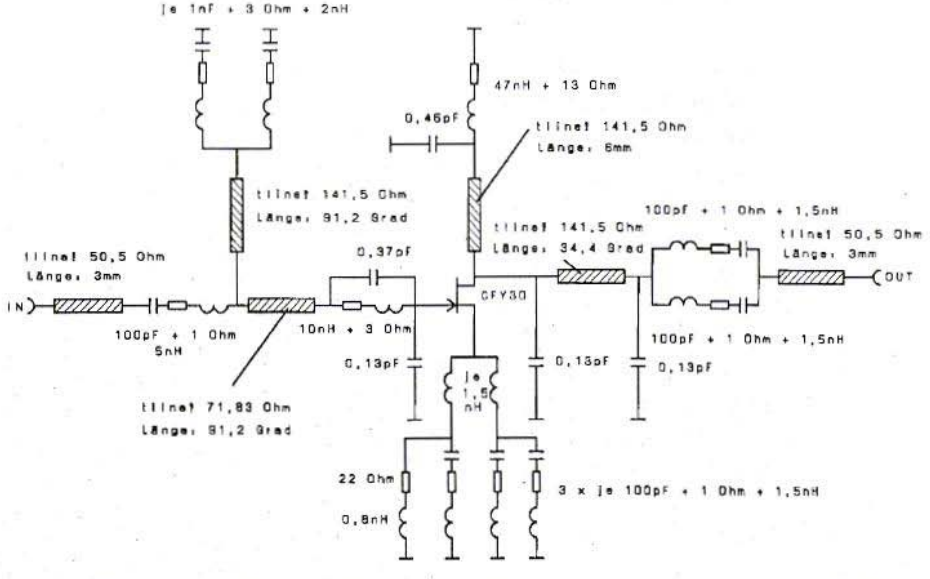

Fig.21: Replacement Wiring Diagram for Circuit with Capacitively Earthed Source Pin
***Länge* = Length**

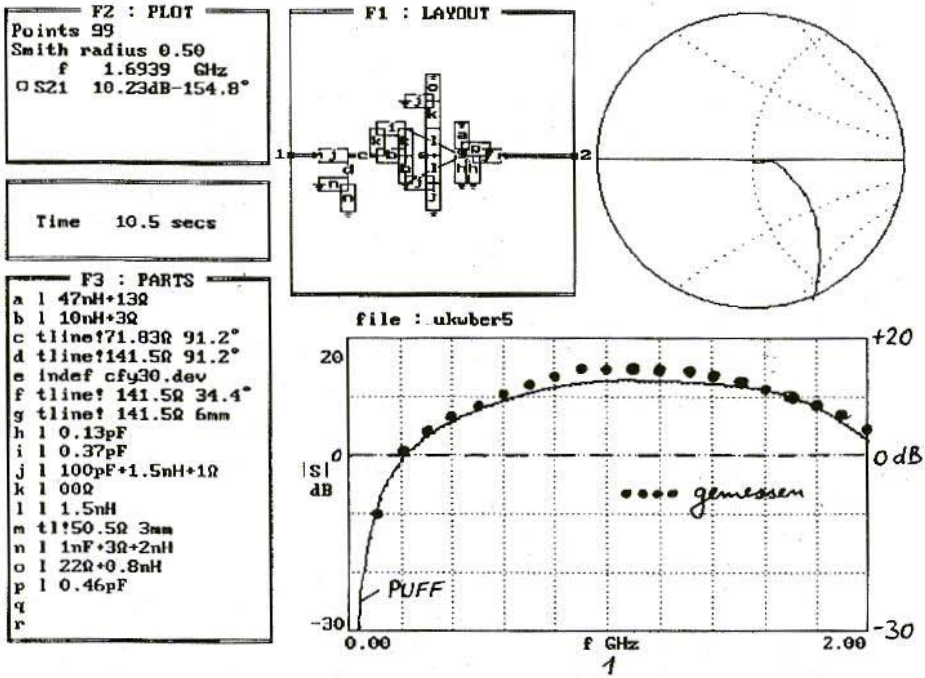

Fig.22: Puff Screen Print-Out: Gain Slope I S21 I for Circuit with Capacitively Earthed Source Pin in Simulation / Measured Values Comparison from 0 to 2 GHz

d) The SMD choke with 47nH (in the drain circuit), has a quality grade of app. 30 to 40. In addition, the unavoidable self-capacitance, of app. 0.2pF, should be taken into account, as this forms the basis for the self-resonant frequency given in the data sheet. Because of its size 1210 SMD, the throttle also needs a soldering scab about twice as big, and with double the capacity.

   For this component, a series circuit is used, with values of 47nH and 13Ω, parallel to which lies a capacitor, with a total value of 0.2pF + 0.13pF + 0.13pF = 0.46pF.

e) At the source connection, the through-platings from the two earth scabs on the top face to the bottom earth surface, in the form of two inductances (each of app. 1.5 nH) between the source pin and the earth potential, are to be taken into account in the simulation.

But in addition the following change is necessary:

The entry for the GaAsFET used in the F3 parts list must be amended from "device cfy30.-dev" to "indef cfy30.dev". That means a "dipole" is suddenly converted into a "tri-polar network", and the source connection, which was previously not accessible, can

now be earthed through the two parallel-wired coils, each of 1.5 nH.

f) On the feed for the negative gate bias voltage, rather high capacity values were set for the earthing using "standard materials" at the end of the quarter-wave transmission line, to make the screening more broad-band. This naturally leads to somewhat higher losses and series inductances, so a new component has to be entered in the F3 list for this.

So we come to the Puff simulation circuit as per Fig.18. Anyone who now starts to complain about the confusion and the lack of room when entering numerous components in layout field F1 may be interested in a tip:

Define a "resistance with 0Ω", i.e. a shunt. You can now extend the "connecting legs" of the components. I have, of course, discovered that this works correctly only at low frequencies, and that in broad-band simulations up to a few GHz - above all, as soon as striplines come into the picture - significant errors will be observed, as against the "original circuit", so be careful!

Finally, Fig.19 shows the new simulation result.

Anyone who is surprised, when the simulation starts, to receive the message "Danger - Circuit too big for Puff, not enough memory" need only make a significant reduction in the number of plot points. In certain circumstances, it makes sense to reduce them to only 50. Sometimes it is also necessary simply to have the main memory organisation of the PC re-optimised, using the DOS service program "Memmaker".

In addition, the readings obtained for the gain slope for the frequency range from 100 MHz to 2 GHz should be plotted into the frequency response curve. The approximation is already very close!

In the range between 1 and 1.5 GHz, Puff prophesies that amplification will be too low, the reasons for which might well be:

a) As the frequency decreases, the board material losses also diminish, but calculations are carried out using a constant loss factor of 0.015, which then becomes too high.

b) For coils and capacitors, Puff can calculate only for a constant loss resistance, but not for a constant quality, as would be the case in

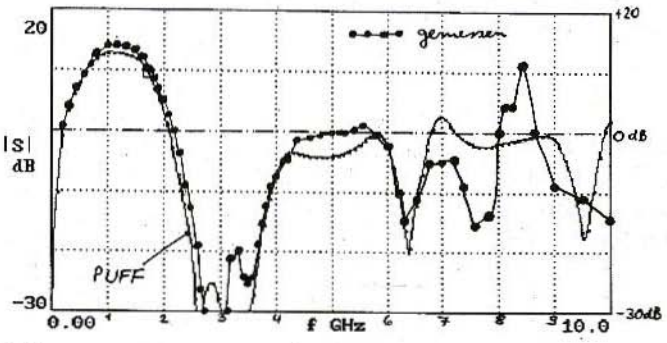

Fig.23: Expanded Simulation Range 0 to 10 GHz: Comparison of and Measured Values Capacitively Earthed Source Pin
***Gemessen* = Measured values**

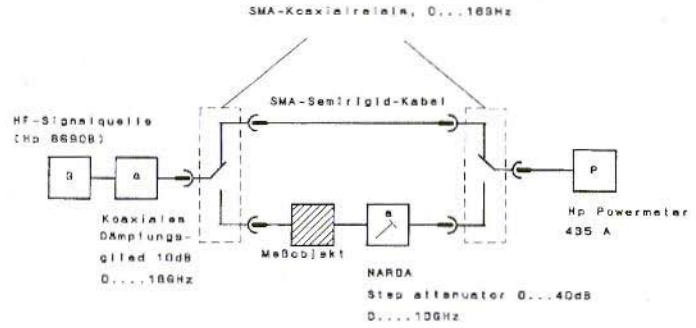

Fig.24: Basic Structure of a Calibration Device
***HF-Signalquelle* = High-frequency signal source, *Koaxiales Dämpfungsglied* = Coaxial damping module, *Koaxialrelais* = Coaxial relay, *Semirigid-Kabel* = Semirigid cable, *Meßobjekt* = Measurement Object**

practice; so the series resistances for blind components are too high at relatively low frequencies

Finally, it was also of interest to know up to what frequency the Puff simulation supplies a result which can be used in practice. So Fig.20 shows the calculated and measured gain slopes in a joint diagram, up to a frequency of 10 GHz.

It can be seen from this that up to about 5 GHz the simulation model drawn up provides results which are confirmed by the measurements. For frequencies above this, a more accurate replacement circuit has to be worked out first, and tests also have to be done to determine whether wave guide effects are already playing a part in the closed housing. Moreover, the capacitive couplings between the various areas of the board (with values markedly below 0.1pF) can no longer be neglected.

6.2.9. Theory and Practice in the Circuit Version with Capacitively Earthed Source Pin

In principle, the existing circuit is used for the simulation, but in the source area it includes the replacement circuits for three parallel-wired capacitors, each of 100pF, and the source resistance with 22Ω (Fig.21). Please remember that even such a little mite of an SMD resistance as a model 0805 has a self-inductance. It was determined by the measurement circuit as approximately 0.8nH and has been immortalised in the F3 parts list. If this is too much, you must help yourself a bit by linking two components in a parallel circuit.

It would also be a rewarding task to reduce the connection soldering scabs for all SMD components used in the high-frequency branch (FET's, throttles, capacitors, resistances, etc.) in order to reduce the interfering capacities towards the continuous earth surface, but the minimum size of the soldering scabs is very firmly based on the solderability

and positioning accuracy of the components!

The rest is now almost routine, and leads to the results depicted in Fig.22 (frequency range 0 to 2 GHz) and Fig.23 (frequency range 0 to 10 GHz). The measured values should again be entered as well, so that they can be compared with the simulation. In the range between 0 and 3 GHz, the response differs only insignificantly from the version with a directly earthed sourced pin. Above this, the chip capacitors begin to cause concern, and at app. 8 GHz create a resonance peak with 12dB amplification.

To sum up, it can be stated that:

a) Right up to 3 GHz and above, the two circuit versions not only display almost identical frequency responses, but display messages which very precisely confirm the predictions made by Puff.

b) If the source connection is earthed through several chip capacitors, then above 3 GHz divergences occur, which have their origin in the impedance cycle of the source earthing component. The circuit with direct earthing of the source connection displays a more uniform response there, without the extreme resonance step-up. For this reason, it should probably be controlled rather more strictly and should display less broad-band noise. You should just test the use of the wickedly expensive 10-GHz ATC chip capacitors, which consist of a solid porcelain chip.

c) For the housing construction, the layout, the selection of components and the mounting, there is a series of rules, the infringement of which is immediately punished by unsatisfactory frequency responses, or even by a tendency to oscillate.

d) As soon as you have some experience in estimating all the effects involved and the behaviour of the components correctly, you can virtually have blind faith in Puff's predictions - to such an extent that, after initially trying out a new circuit with Puff, you can draft more and more complicated replacement images, to determine where measurement divergences will arise.

6.3. A Few More Measuring Tips

To conclude the chapter on amplifiers, a few more hints on measurement technique and on plotting frequency response curves:

The ideal, and most modern, method, is to use a network analyser. Fortunately, there are fellow radio amateurs who use this kind of equipment at work, and can occasionally run something through. However, in order to reach one's goal with one's own measuring equipment, which is usually rather out-of-date, two things are necessary:

a) Increased patience and time for "manual measurement", since this involves rotating, adjusting, recording and correcting, and:

b) Constructing a measurement apparatus which imitates the "self-calibration process" of modern equipment (Fig.24). Here, you can simply

switch from "Calibrate" (= signal passing through a section of cable) to "Measure" (= signal passing through the object of measurement), or vice versa, with the help of two SMD coaxial relays. If, in the "Calibrate" position, we now turn the output voltage up until the power meter shows precisely end-scale deflection, "Zero dB", we then finally just have to switch over to "Measure" and observe the rise or fall in the instrument reading.

The change in the display then exactly matches the desired amplification or damping in dB.

Many inherent errors and many frequency responses from the measurement equipment, cables and connecting plugs in use are eliminated in this way, as they have the same effect in the calibration channel and the measurement channel, and we are interested only in the differences.

Of course, different transmission loss readings from the two relays lead to an incorrigible and permanent error, so only the finest kit must be used for this (military versions, gold-plated SMA sockets with a frequency range up to 18 GHz, and crosstalk attenuation of min. 60dB, reflection factors below 10%, etc.). In the same way, you must fake an ideal moving load to the test transmitter by means of a subsequently switched-in coaxial attenuation member, to keep down the level of any additional measurement errors caused by a poor input reflection factor for the object of measurement.

Since, in addition, the measurement range of the power meter does not usually exceed 40dB, a reversible calibrating line, with 0 to 40dB and a usable frequency range going up to over 10 GHz must be positioned after the object of measurement if it is, for example, an amplifier. In this way, the rise in the level after an amplifier can be damped down until it is again within the range of measurement of the power meter.

(To be continued)

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Gunthard Kraus, DG8GB

Design and Realisation of Microwave Circuits Part 4

7. MMIC AMPLIFIER

7.1. Preliminary Observations

At two points in our "Meteosat Converter" application project, we need at least 20dB amplification with simultaneous tolerable noise background:

a. Directly behind the "low-noise preamplifier" referred to in Section 6.2. of Part-, for the significant increasing of the signal level before the subsequent 1,700 MHz band pass, with simultaneous damping of the passive mixer.

b. Directly behind the mixer, or in the downstream intermediate-frequency stages to have a signal level sufficient for a relatively long coaxial cable connection from the converter (near the antenna) to the evaluation circuit in the building.

The use of a "universal amplifier module" from 50 Ohm technology is on offer here; to this end, MMIC's (microwave monolithic integrated circuits) are available in the most varied formats, and with varying limiting frequencies.

After a look through the relevant literature with application circuits, the choice finally fell on type INA 03814 from Avantek/HP, because of the current price and delivery situation.

We find the following specification for this MMIC in the data sheet; 25dB amplification in the range up to 2 GHz, for noise factors of about 3dB.

The S-parameter can also be found, on the Avantek data diskette, which can be supplied for Puff.

Here you merely need to convert the "touchstone file" (INA03184.S2P) into Puff format (which gives INA03184.dev). As mentioned, the conversion program you need for this is supplied with the Puff diskette.

But don't forget to estimate the values for the "zero Hertz" frequency and add them to the *.dev file!

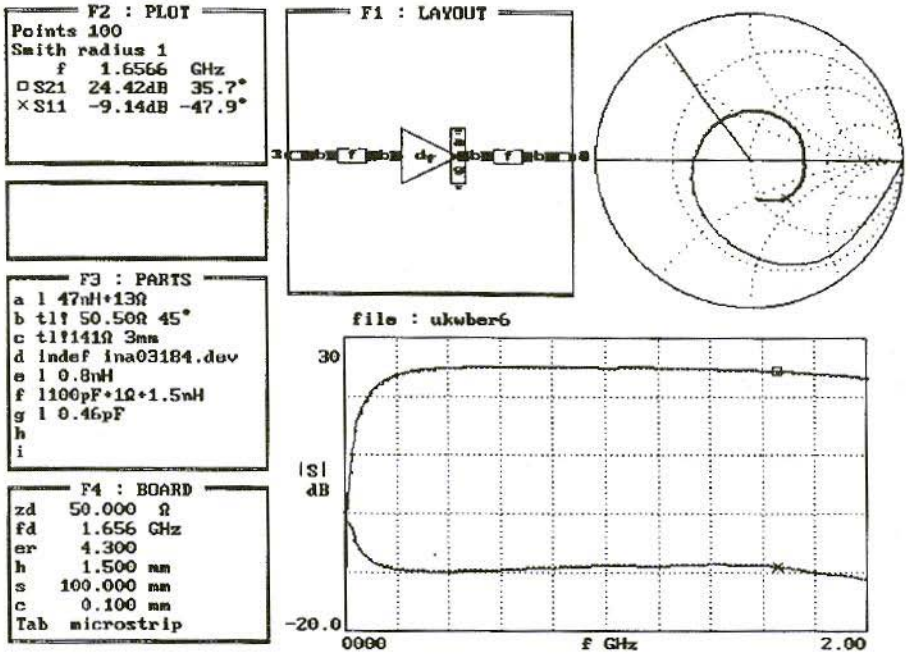


Fig.25: Ideal MMIC amplifier with INA 03185

7.2. Draft Circuit and Simulation

To judge from what many semi-conductor manufacturers say, drafting a microwave amplifier circuit is child's play, and is exceeded in simplicity and operational reliability only by the drafting of a power supply circuit with a half-wave rectifier.

The input and output pins are simply connected using 50 Ohm microstrips, into which are inserted the DC blocking capacitors. The power supply DC voltage is fed in through a 47nH SMD choke.

If, during the simulation, the same equivalent circuits are used for chokes and capacitors as for the low-noise preamplifier, you obtain a wonderfully smooth amplification curve between 1 and 2 GHz, with a peak value of app. 27dB. Fig.25 shows the MMIC amplifier simulation.

However, in reality the assembled circuit will initially oscillate at f = 1.4 GHz. Moreover, it is almost impossible to produce any attenuation by means of the well-known "calming measures" - e.g. using conducting foamed material, etc..

If we now make use of the experience gained in the drafting of the LNA and test the influence of the "earth feed-throughs", using Puff, we discover something really amazing:

Even a total inductance of 0.3nH, applied for the feed-through of the two earth surfaces of the IC to the bottom earth plane, is sufficient to make the

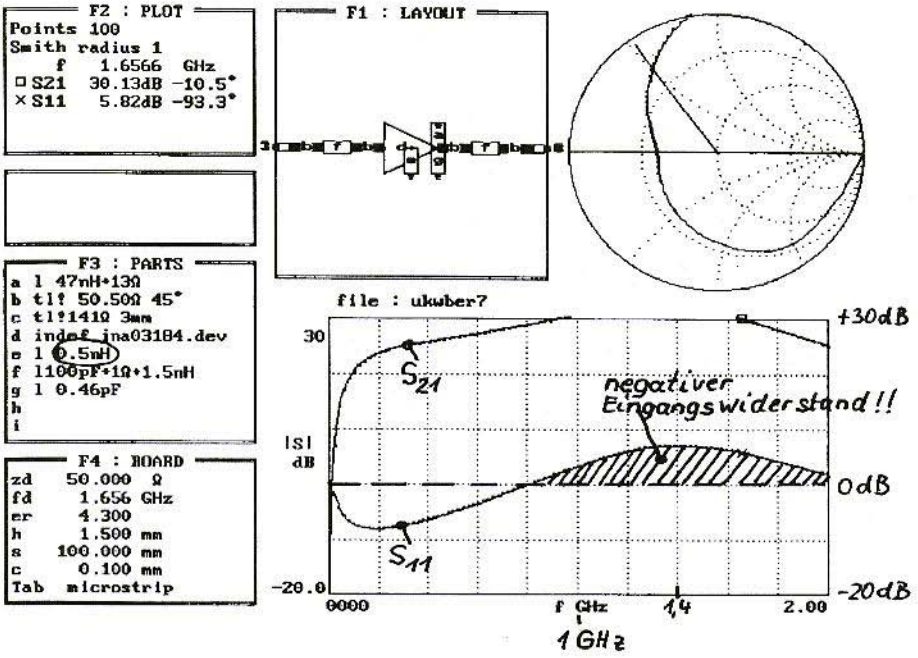


Fig.26: Simulation with Puff: MMIC amplifier with L = 0.5nH in earth branch

input reflection factor in the frequency range at 1.4 GHz markedly greater than 0dB (Fig.26 shows the ratios for 0.5nH).

The circuit then has a negative input resistance. In these circumstances, it can't help but oscillate!

We must therefore use all means to ensure that this inductance falls; Fig.27 shows how this is done.

a. A hole with the external diameter of the MMIC housing is drilled in the printed circuit board.

b. The top earth surfaces are now fed as a wedge into the layout until they are directly against the housing of the MMIC.

c. The feed-throughs are started as close as possible to the housing. 5 holes are drilled on each side.

It can probably be easily recognised from the diagram that the question of the insulated earth surfaces for input, earthing and output has again been taken to heart (it's just that putting in the many 0.8 mm. rivets, with a "watchmaker's power riveting machine", specially created for this, as an extra, that really is a brute of a job!)

The output and input microstrips require special consideration at 50 Ohms. They should have a "wedge-shaped point" , and then be reduced, right down to the width of the IC terminal lug, thus avoiding additional inductances and points of impact.

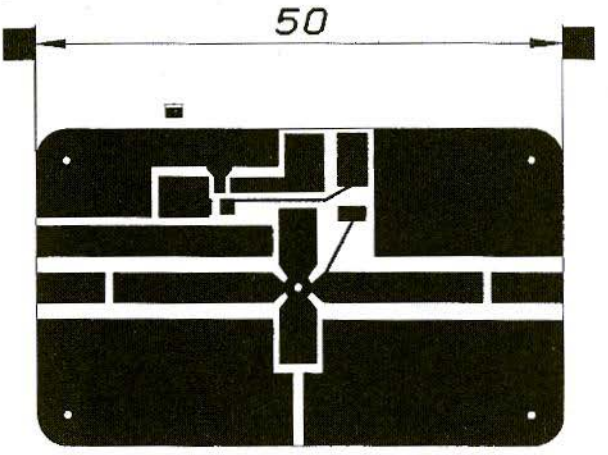


Fig.27:
Layout of MMIC amplifier for 1.7 GHz

The already well-known 47nH choke for feeding in the operating voltage is here too as well, together with its 6mm. long feed and its joint built-in capacitance of 0.46pF and the 13 Ohm losses in the equivalent circuit diagram.

The coupling capacitors in the input and output circuits are laid out, as usual, as two 0805-NP0 types in parallel, at 100pF. Firstly, this precisely fills out the width of the microstrips and thus reduces the reflections, and secondly, the

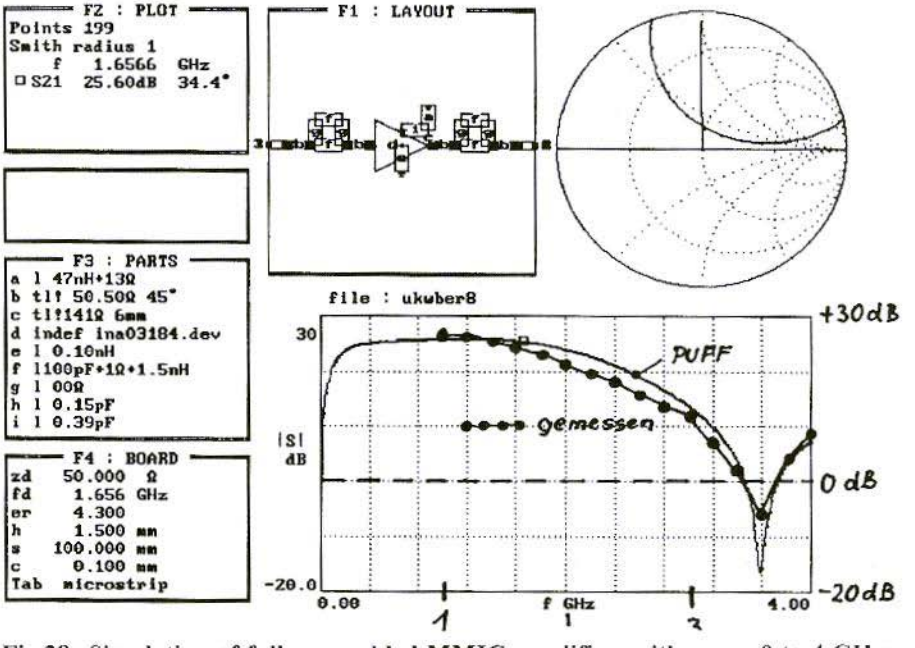


Fig.28: Simulation of fully assembled MMIC amplifier, with range 0 to 4 GHz; *Gemessen* = Measured

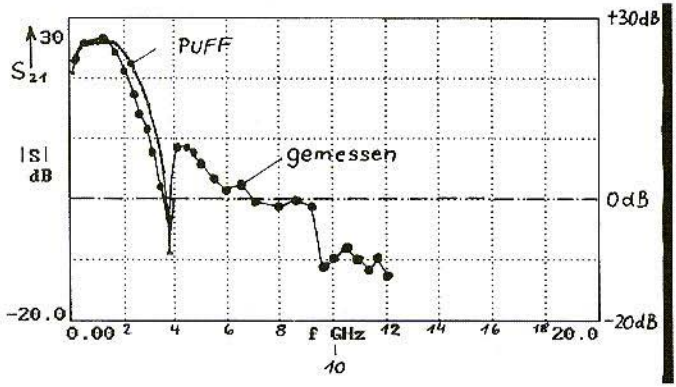


Fig.29: MMIC amplifier: gain slope |S21| from 0 to 12 GHz *Gemessen* = Measured

joint inductance of 1.5nH and the serial loss resistance of 1 Ohm are halved by using one component.

Fig.28 displays the full results of the simulation, including the measured results for up to 4 GHz. Of course, immediately after the measurement the simulation was adjusted to the reality.

The onset of amplification at 3.6 GHz is actually brought about by the 6mm long microstrip and the 47nH choke, with the (assumed) 0.46pF capacitor! Here a short-circuit is obviously transformed to the amplifier output, and we must reduce the 0.46pF to 0.39pF for the Puff simulation to supply the same resonance frequency. There is a quite simple reason why the minimum value in the actual assembled circuit is not as low as in the prediction. It is well-known that the losses for FR4 equipment increase more and more sharply from 1.5 to 2 GHz, and here we are always stimulating with a constant loss factor of 0.015. The assumed quality is therefore too high, and there is a discrepancy between reality and the Puff result.

Anyone who would like to have a more precise result would first have to measure the valid loss factor for this frequency, and then simulate for only a narrowly restricted frequency range for which this determined factor is approximately constant. For the same reason, the measured amplification in the range between 1.5 and 3.4 GHz is probably always somewhat lower than the theoretical progression, since - because of the poorer quality - as in every resonant circuit, the resonance curve of this "resonant circuit" must be flatter and broader.

One more little thing:

The S parameters of the INA 03184 are specified by the manufacturer only up to a maximum frequency of 4 GHz. But if, for example, an upper plot limit of 10 GHz is inadvertently entered, a bleeping noise is suddenly heard when the 4 GHz limit is exceeded, which in normal circumstances would be the acoustic signal for a computer fault. But the program has not crashed. It continues operating and bleeping, right up to the highest frequency entered, and you need only wait for the computer to quieten down.

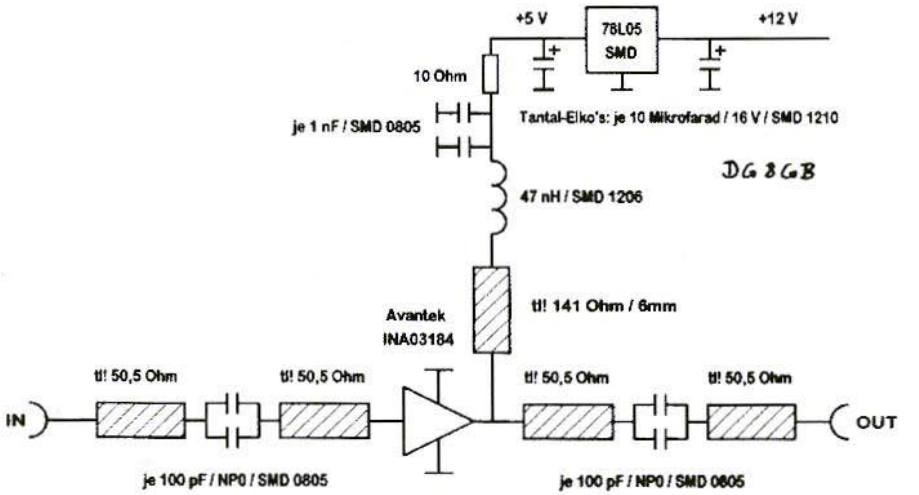
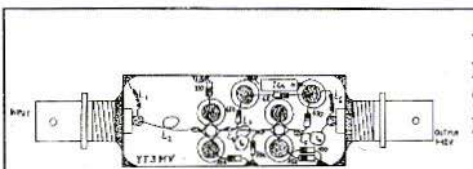


Fig.30: Wiring diagram of MMIC amplifier prepared for Puff;
***Tantal-Elko's* = Tantalum Electrolytic Capacitors,**
***Mikrofarad* = Microfarads, *Je* = Every**

The diagram created using a network analyser (HP 8410A) shows the simulation behaviour of the circuit up to 12 GHz.

Finally, Fig.30 shows the wiring diagram for the layout, which can certainly be analysed without problems.

(To be continued)



Very low noise aerial amplifier for the L-band as per the YT3MV article on page 90 of VHF Communications 2/92. Kit complete with housing Art No. 6358 £36.55. Orders to KM Publications at the address shown on the inside cover, or to UKW-Berichte direct. Price includes p&p

Gunthard Kraus, DG8GB

Design and Realisation of Microwave Circuits Part 5

8. THE STRIPLINE BAND PASS for 1,700 MHz

8.1. Preliminary observations

Behind the pre-amplifier chain there follows a band pass, which is to ensure that only the input frequency range around 1,700 MHz is made available to the mixer. It also prevents the oscillator frequency effective in the mixer from seeking a way back to the antenna through the amplifier.

So we must lay down the following requirements:

a) Insertion attenuation as low as possible for the Meteosat reception range around 1.700 MHz

b) Insertion attenuation as high as possible for the oscillator frequency (used in this design) around 1,600 MHz

c) Insertion attenuation just as high for the image frequency range of 1,500 MHz, to prevent the noise there from being transferred onto the intermediate frequency selected here (app. 100 MHz) and thus making the noise factor of the entire system worse.

That means a pass range curve "displaced right", which still lets 1,700 MHz through well (just), but on which the oscillator frequency and the image frequency are already as far out on the "sloping flanks" as possible, and are strongly attenuated.

The question of the type of filter assembly now arises. Among the various options for the gigaHertz range: helix filter, inter-digital filter, Stripline filter, coaxial filter, etc., the Stripline filter was deliberately picked out, for the following reasons:

a) To gain experience in the creation of such filters - both in simulation using Puff and in the production of the printed circuit

b) Inexpensive manufacture

c) No balancing of any sort required

d) Absolute reliability for copying, if the

scattering in the track dimensions or the board material properties is kept small in series manufacture.

The relatively large dimensions should be pointed out as the biggest disadvantage of this filter as soon as the degree of filtration is increased in order to obtain steeper flanks. The only way out is to use board material with very high dielectric constants and correspondingly large shortening factors - for example, RT-Duroid 6810 would be normal here, with e_r = 10.4 to 10.8.

8.2. Dimensioning documents for Stripline band passes

Unfortunately, Stripline band passes are not as easy to calculate and dimension as a parallel oscillation circuit. What we are actually dealing with here is a series circuit of "coupled line sections, each with 90 degrees of electrical length", with the high computing power of modern computers being indispensable for the analysis and dimensioning programs which are required.

Here there are only two options available to the user in normal cases:

a) To use the third-degree "bpf.dev" specimen deep pass supplied on the Puff diskette - i.e. with three consecutive, coupled microstrip line pairs - and to dimension it for the board material used, the band pass frequency provided for and, as far as possible, the required band width.

b) To obtain relevant literature on this topic, struggle through it, and then design the appropriate filter yourself, but now at a markedly higher degree of efficiency, to obtain a better characteristic.

Once again, both paths were followed, so that we could obtain a direct comparison between the necessary expenditure on a specialist literature search and on Puff simulation, and obtain the resulting readings for the specimen circuits.

8.2.1. Working with "bpf.puf" specimen band pass

Call up the "bpf" file and enter the medium frequency, 1,693 MHz, the dielectric constant (most recent version: 4.32 at 1,693 MHz) and the board thickness, 1.5 mm. in field F4.

Then carry out an initial computing cycle.

Default in F2: 200 computation points, frequency range 1-2 GHz, value range for S_{21} and S_{11} from 0 to - 25 dB. Finally, press "p" for "plot".

First store the result under a suitable name - for example, BP1700M - and leave Puff for a short time.

Now call up the word processing program and load file "BP1700M". Now enter the data valid for the FR4 material as already used:

Dielectric constant, ε_r = 4.32
Dielectric loss factor, lt = 0.015
Thickness of copper, mt = 35µm
Surface roughness, sr = 0.002 mm.
Minimum conductor spacing = 0.2 mm.

Store the file again (without word processing formatting signs!), re-load Puff, call up file "BP1700M" again. Insert the now familiar exclamation

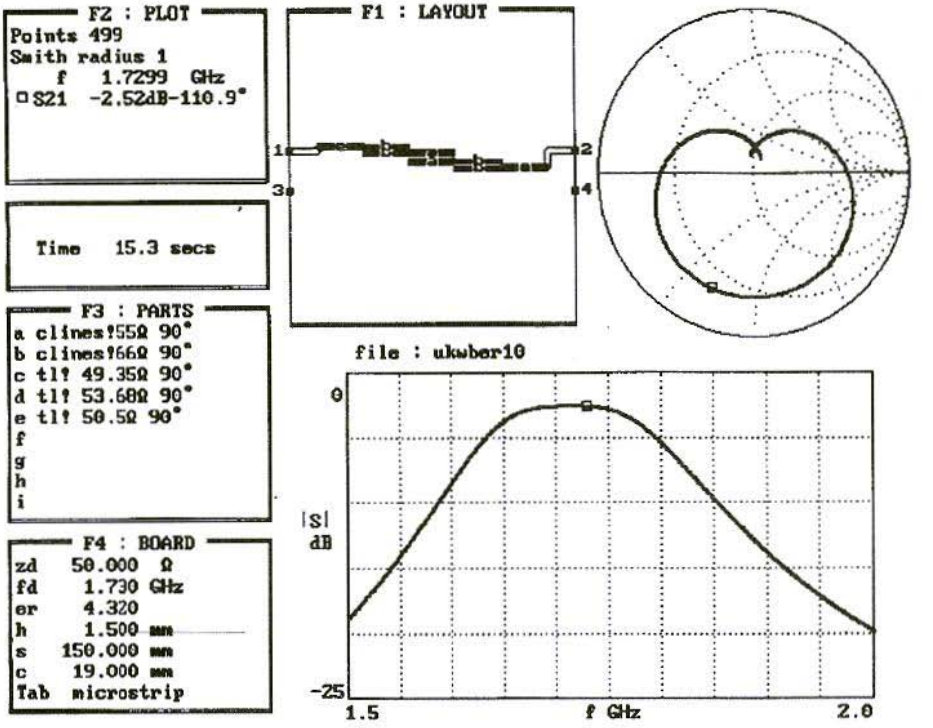


Fig.31: Simulation of Lossy Bandpass

mark in field F3 for each Stripline, behind the word "clines" (which can be abbreviated to "c" without any problem).

Now start another plotting procedure, to obtain a realistic prediction of the pass range curve and transmission loss the completed filter will display when engraved.

If the medium frequency has to be displaced further, this can be done in field F4, under item "fd", while the band width, and thus the form of the pass range curve, can be changed by altering the resistance values in the two coupled Striplines in the components list, F3.

The following conditions apply here:

a) If the impedance level given in F3 (this is the "Even" value) for one of the two line pairs increases, the band width increases. Of course, as in a two-circuit band filter, the curve takes on the typical "dip" caused by over-critical coupling.

But be careful. Changes of 0.5Ω are already sufficient to produce a sharp bend in the curve!

b) If the resistance values are reduced, the "under-critical coupling" case is obtained. The pass range curve becomes narrower and narrower, until suddenly the transmission loss increases sharply and the reflection factor becomes worse.

In Fig.31, we can see the simulation for the filter afterwards converted into practice, with two 50Ω Striplines added as a link to the SMA sockets at the input and output.

Another important tip concerning Puff. How do we get the individual line pairs together for a Stripline filter?

Anyone who, for example, has changed or deleted too much in the layout in window F1 would certainly prefer to re-draw the circuit.

So first we delete everything, using <Control e> as usual. Then we desperately attempt to combine the individual line pairs from list F3 together into a Stripline band pass again. The trick here is as follows:

Position the cursor as far to the left as possible in field F1, lay down the 50Ω stripline and connect the input with pin-1 of the layout by pressing "1". Then use the appropriate letter to select the first line pair, and then press the "Cursor Right" key. Now simply select the next line pair, one after the other, and keep pressing "Cursor Down". So you go all the way to the end again, and finally you can connect the output with connection "2" through a further 50Ω circuit.

Unfortunately, this procedure is not documented in the manual!

Now we just need the mechanical dimensions of the Striplines for the board design using a CAD system. Anyone who briefly deletes the call-up sign for each line and keys in the equals sign obtains the following details:

Line a:
(Ze = 55Ω), length = 23.96mm.
Width = 2.984mm, interval = 1.696mm

Line b:
(Ze = 66Ω), length = 24.385mm.
Width = 2.593mm, interval = 0.281mm.

To this must be added a width of 2.88mm for the 50Ω supply line.

But be careful! The lengths need to be corrected again, because at the end of all open lines the electrical field lines go on a little bit further. But this increases the electrical length of the line! Here this means that the line sections in the board layout must be mechanically shortened so that the resonance frequency is correct again.

There are two different options for doing this as well.

8.2.2. Length correction for Striplines

Case A)

The Stripline ends abruptly and is not followed by another link.

For this case, you can find a diagram in the Puff manual (p. 38) for determining the shortening, which is repeated as Fig.32. Of course there is some more work for Puff first, because we must first use samples to determine what the impedance level of each individual line section is, with its special conductor width.

This is not particularly difficult. You simply enter a "Normal Microstrip Line (tline)" in components list F3, with any length, which has an impedance level of app. 50Ω. Press the "Equals" key and you will obtain the associated conductor

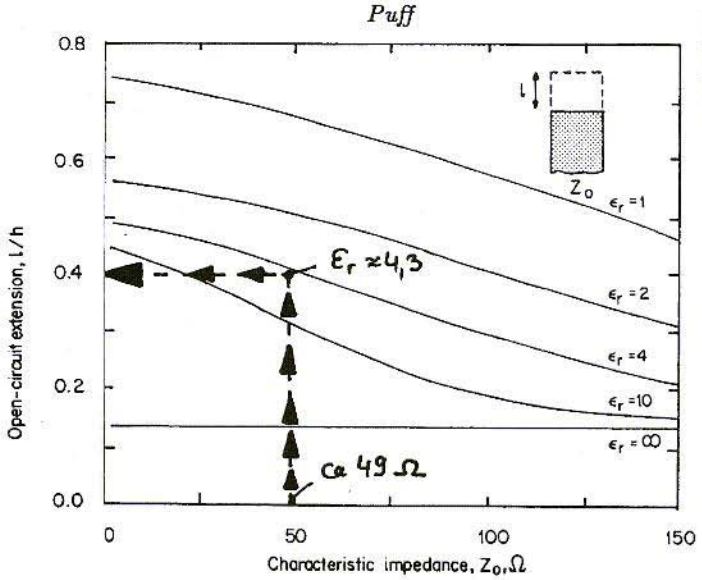


Fig.32: Determination of Electrical Shortening

width. Now keep changing the impedance level until this conductor width is identical with that of the individual track of the coupled Stripline.

With the dielectric constant, 4.32, and the impedance level of the line section, you can now go into the diagram and read off the "Open End Shortening" required as a fraction of the board thickness.

You will obtain the following results for the two line pairs in question:

Line "a":

The track width is 2.984mm., which corresponds to a lossy Microstrip line (= "tline! 49.35Ω, 90°"), but which would have an actual impedance level of 48,93 Ω. For this, in combination with the dielectric constant of 4.32, the diagram gives us an electrical extension which is equal to 0.4 x board thickness = 0.4 x 1.5mm. = 0.6mm..

Line "b":

A track width of 2.593mm. gives us a "tline! 53.68Ω, 90°" with an actual impedance level of 53.154Ω for the individual track.

For this, the diagram gives an electrical extension of 0.39 x board thickness = 0.39 x 1.5mm. = 0.58mm..

And now for something I came up against when measuring out the first board I set up, which really made me realise the discrepancies between the measured results and the Puff simulation (app. 2%):

The extensions obtained are valid for "simple" Striplines. Here the result agrees precisely with the theory. However, for the coupled lines we are talking about here there is always an inductive coupling between the two tracks.

Obviously, if the electrical field lines extend a little further at the ends, this

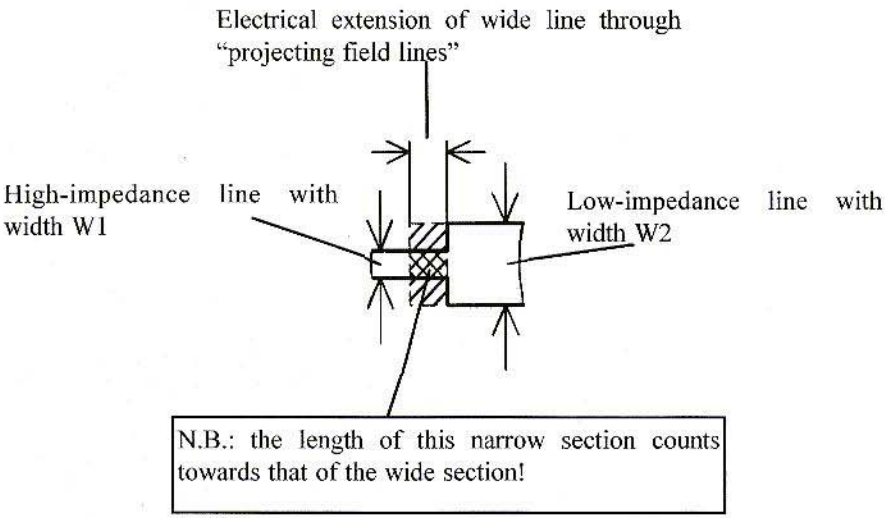

Electrical extension of wide line through "projecting field lines"
High-impedance line with width W1
Low-impedance line with width W2
N.B.: the length of this narrow section counts towards that of the wide section!

Fig.33: Representation of Electrical Extension

decreases more or less linearly down to zero, as soon as we (in imagination) leave the track, when no more copper is present and the current can no longer flow in it.

For this reason we obtain the correct resonance frequency only if, as an exception, we do not take the entire extension into account, but consider only the half of it on either side of the line section, i.e. app. 0.15mm. in each case.

Case B)

An additional, narrower line is connected to a wide Stripline.

Since the final capacity of the wide line is now reduced by the copper track on the subsequent narrow track, the electrical extension is also reduced. In addition, the cross-hatched "narrow extension section" is always added onto the wide line and omitted from the length of the narrow line. This loss of length must then be appropriately taken into account in the layout of the narrow line and must be balanced out (see Fig.33).

The correction formula for a "discontinuity" of this kind runs as follows:

$$X_{Separate} = X_{wide} [1 - (W_1/W_2)]$$

where

$X_{Separate}$ = Actual effective extension at discontinuity

X_{wide} = Theoretical extension of wide line alone (from top diagram in Fig.32)

W_1 = Conductor width of narrow Stripline

W_2 = Conductor width of wide Stripline

We then obtain the following correction values for the "discontinuities" for the Stripline band pass:

Conductor lengths as per Puff

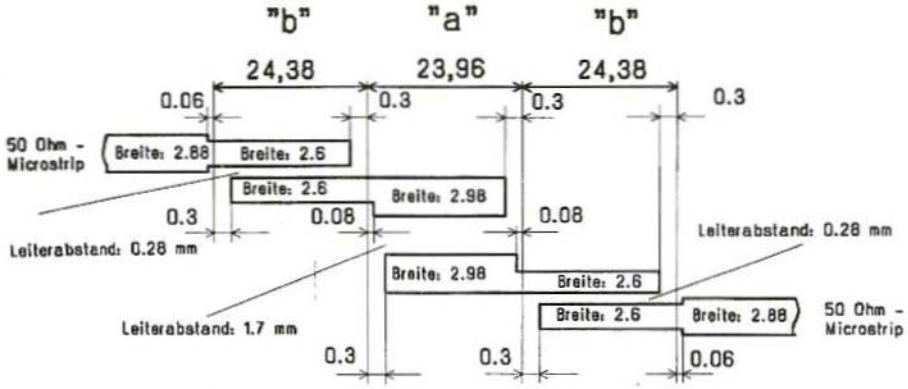


Fig.34: Difference between "mechanically" and electrically effective Conductor Lengths
***Leiterabstand* = Conductor interval, *Breite* = Width**

1. Transition from 50Ω supply line (width 2.88mm.) to coupled line "b" (width app. 2.6mm., associated impedance level of a single line would be 53.153Ω):

Shortening = 0.6mm. x (1 - 2.6mm. / 2.88mm.) = 0.058mm., i.e. 0.06mm.

2. Transition from line "b" (width 2.6mm., the corresponding impedance level would be 53.153Ω) to line "a" (width 2.98mm., corresponding impedance level 49.35Ω):

Shortening = 0.6mm. x (1 - 2.6mm. / 2.98mm.) = 0.0765mm., i.e. 0.08mm.

Now carefully enter all this into the board layout using CAD. Fig.34 gives a block diagram of the points to watch out for in designing the board. You can now clearly see the difference everywhere between the "mechanical" lengths to be

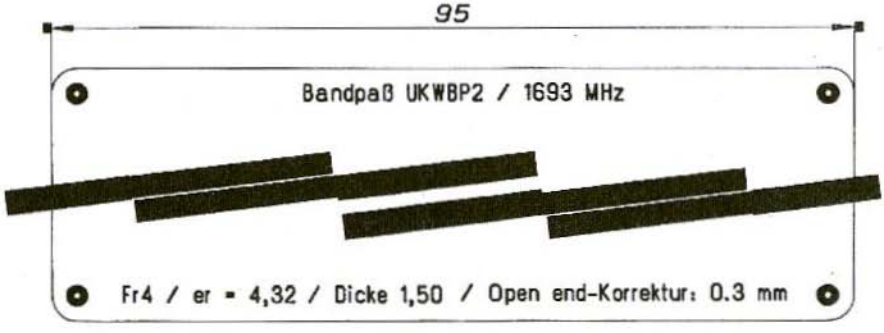


Fig.35: Completed Board Layout of Band Pass
***Bandpaß* = Bandpass, *Dicke* = Thickness, *Korrektur* = Correction**

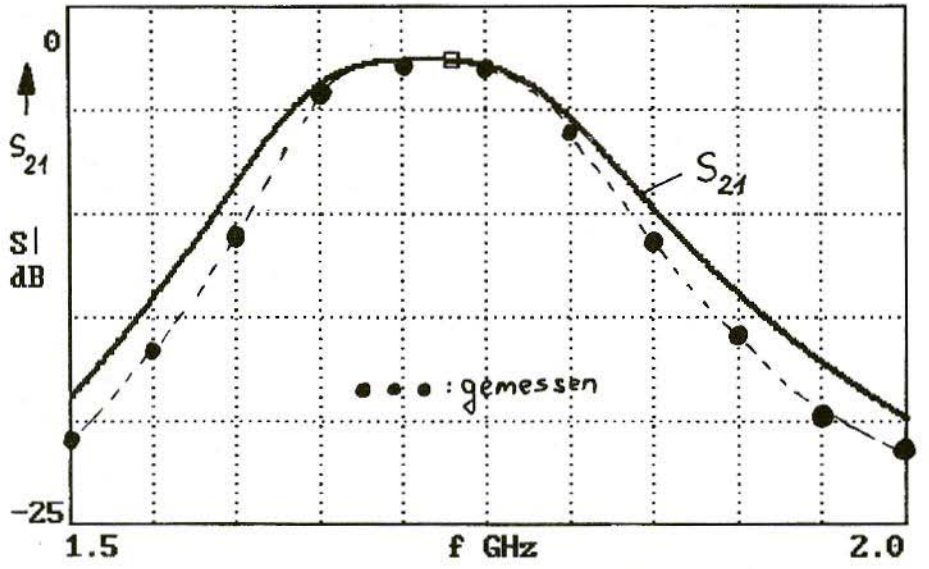


Fig.36: _S21_ of Band Pass for Range 1.5 to 2 GHz; *Gemessen* = Measured

printed and the electrically effective lengths.

Naturally this involves a lot of effort and a lot of computing time, but this is precisely where what we are doing differs from the very expensive professional programs referred to, which supply the complete layout at the same time as they do the computing.

8.2.3. Instructions for assembly

Because of the small conductor interval in the coupled line "b", the layout result absolutely must be printed out highly magnified, and then reduced again photographically when a film is made. Only in this way can "short circuits" be avoided, since photographic copying is far more precise than a printer. Naturally things are even simpler and even more precise with a photo plotter.

Fig.35 finally shows the completed board layout.

Things to look out for here: the 50Ω connection lines at the input and output were carried on precisely along the extension of the band pass line pair, in order to avoid a bend. In this way, the computed shortening values continue to correlate, whilst in addition any reflection errors arising due to the "oblique" connection of the SMA plug can be compensated for relatively easily if necessary.

In Fig.36, the computed and measured pass range curves in the range between 1.5 and 2 GHz can be compared with one another.

As can be seen, our observations were pretty well correct. Something which seems rather puzzling at first is the fact that the flanks fall away rather more steeply on the measured board than in

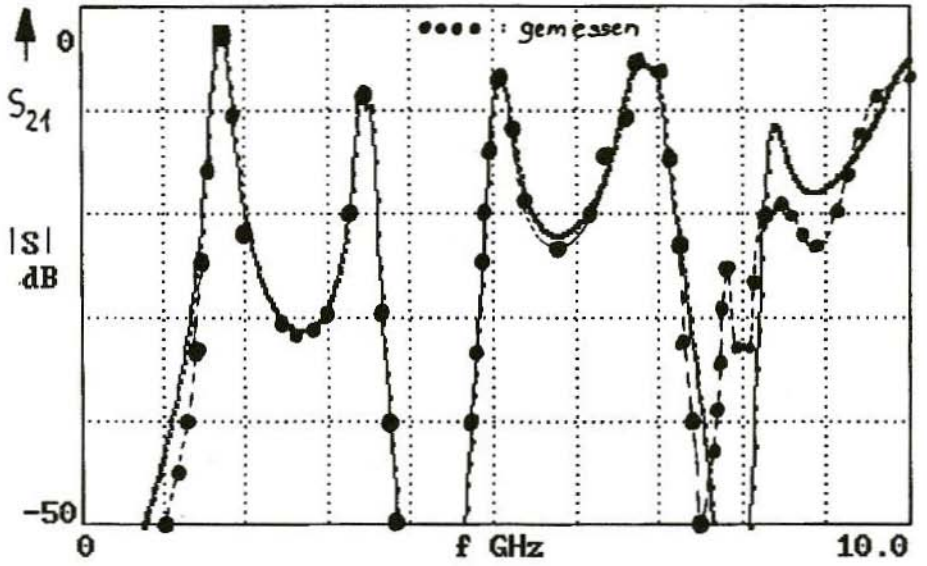


Fig.37: _S21_ of Band Pass for 0 to 10 GHz; *Gemessen* = Measured

the simulation, for clearly the "band filter coupling" is not quite as firm as it was computed to be.

As one possible explanation, when I examined the board with a magnifying glass (!) I found something which one would hardly have suspected in advance.

The resolution of the laser printer used is only 300dpi, which leads to the notorious "stair nosing" on the tracks of the Stripline running at an angle to the board axis in the board layout. Obviously, with the very narrow interval between the two "b" lines, this is already enough to make the coupling looser than was foreseen, and thus the pass range curve has steeper flanks. In addition, an investigation ought to be carried out, using professional methods, to determine the extent to which any "under-cutting of track edges" is also a contributing factor.

The housing construction also has a strong influence, in that if the screening cover is screwed off, then the band width expands markedly and the pass range curves measured match the simulation considerably better!

In Fig.37, we see again the curve from 0 to 10 GHz. Here there is now really no cause for complaint, as there is almost no difference between measurement and simulation, apart from a narrow area around 8 GHz.

Incidentally, the recurring "intrusions" of the attenuation in the filter attenuation band with this filter construction are unfortunately inherent in the principle involved, and are described as "scatter resonances".

If they cause interference, you just combine the above filter with an additional Stripline deep pass on each side, which forms just as good a barrier there!

This two-sided "framing" of the band pass with two equal deep passes has already been shown to be correct, and has become a standard technique in industry, since it keeps the layout symmetrical.

For this application of course, Puff reaches its limits with the first calculations, and suddenly feels faint. Perhaps you remember seeing the message come up - "CIRCUIT TOO BIG FOR PUFF".

In such a case it is recommended that you transcribe the complete band pass-*.puf file to a device file with the termination *, as already done for the GaAsFET CFY30, with the text processing program, and thus simply make a component out of it, which is managed in the F3 list (e.g. as "BP1700M.dev") for further calculations.

Of course, we must not forget to plot the frequency response of the band pass from 0 to 10 GHz before setting up the *.dev file, so that later we will not be surprised by the message that the permissible frequency range has been exceeded.

Moreover, when plotting, use the maximum possible 499 plot points, as otherwise the entire curve looks very stepped if additional deep passes with smaller frequency ranges are analysed later on.

8.3. General calculation of Stripline band passes

8.3.1. Preliminary comments

To work myself into this topic properly, I carried out some research into the relevant literature first.

The most important literature references are, for example, the "original article" on this topic by S.B.Cohn [1], with the corresponding calculation mechanisms, or the further development of these thoughts in the article by G.L.Matthaei [2]. And in German there is a chapter with all the calculation formulae, together with an example showing the working out, in the book by Geschwinde & Krank [3]. Finally, a modern work, which supplies answers to most questions on passive components in microwave engineering, and which no enthusiastic designer should be without. It is distinguished by the way it explains things, which is laudably educational, and at the same time mathematically correct, but very clear. It also includes examples of calculations, not only for this topic, but for nearly all the questions dealt with, plus a very comprehensive list of literature references: Peter A.Rizzi [4].

8.3.2.1. Determination of general filter coefficients for a pre-set reflection factor

The starting point for all filter calculations is the so-called "standard deep pass" for a specific degree of filtration, which simultaneously gives the number of components required.

From this deep pass circuit, by means of appropriate "transformations", all other filter formats can be created, such as high pass, band pass or band elimination, as well as any characteristic desired, such as impedance level, limiting frequency or band width.

Before starting the design work, you

should devote some thought to the following points:

a) How steep should the transition of the filter curve from the pass range into the filter attenuation band be? The degree of filtration, n, depends on this, and thus the number of components or assemblies. A higher degree of filtration gives a steeper gradient. In this context, you should also know that the degree of filtration should, as far as possible, be odd (i.e. n = 3, 5, 7, etc.). The reason is that, if n = 2, 4, 6, etc., the terminal resistance of the filter is no longer equal to the internal resistance on the input side, and also changes with the filter attenuation selected. In many cases, this is very impractical.

b) Should the attenuation and group delay time in the pass range oscillate within specific limits, or must they be constant, as far as possible? In this connection it is always true that smaller oscillations in the "pass range" unfortunately simultaneously create a "flatter rise" in the filter attenuation outside the pass range.

c) The maximum attenuation value in the pass range is always associated with a reflection factor and / or a specific maximum voltage standing wave ratio (VSWR), which can be precisely calculated. The filter calculations can not be carried out until it has been determined.

Fig.38 shows such a standard deep pass with its filter coefficients, g1 to g5, the two assembly options - anti-capacitance or anti-coil - and the essential cycle of the transmission loss for the two most important types of filter, Butterworth and Tschebyscheff. "Butterworth" gives a lovely smooth curve with slight distortions, but in practice the Tschebyscheff filter is often preferred, with its steeper flanks, and in return we accept the attenuation "ripple" in the pass range.

A)

Immediately below, we give the formulae for determining the coefficients for Butterworth deep passes.

The following generally applies here:

where

$$g_k = 2 \cdot \sin\left[\frac{(2k-1)\cdot\pi}{2n}\right] \qquad (1)$$

k = 1, 2, 3,...n

Example for n = 3:

$$g_1 = 2 \cdot \sin\left[\frac{(2\cdot 1-1)\cdot\pi}{2\cdot 3}\right] = 1{,}00 = g_3$$

$$g_2 = 2 \cdot \sin\left[\frac{(2\cdot 2-1)\cdot\pi}{2\cdot 3}\right] = 2{,}00$$

B)

Things are unfortunately rather more difficult with the Tschebyscheff type:

First step:

Determination of maximum attenuation in pass range from reflection factor and/or VSWR.

Conversion of VSWR into reflection factor

$$r = (VSWR-1) / (VSWR+1)$$

The maximum attenuation in dB in (2) pass range then amounts to

Example:

$$a_{max} = 10 \cdot \log \frac{1}{1-|r|^2} \qquad (3)$$

A reflection factor of r = 10% means that

Sec-

$$a_{max} = 10 \cdot \log \frac{1}{1-(0{,}1)^2}\,dB = 0{,}043648\,dB$$

ond step:

We now need the variable ß:

$$\beta = \ln\left(\coth \frac{a_{max}}{17{,}37}\right) \quad (4)$$

For a reflection factor of 10%:

$$\beta = \ln\left(\coth \frac{0{,}043648}{17{,}37}\right) = 5{,}98634$$

We can thus determine the next variable, γ:

$$\gamma = \sin h\left(\frac{\beta}{2n}\right) \quad (5)$$

As an example, we shall now go through all the calculations and the assembly for a fifth-degree "specimen deep pass" (n = 5) with the reflection factor r = 10%.

This gives us

$$\gamma = \sin h\left(\frac{5{,}98634}{2 \cdot 5}\right) = 0{,}635$$

Third step:

The individual filter coefficients are determined according to the following formula:

$$g_1 = \frac{2 \cdot a_1}{\gamma} \quad (6)$$

$$g_k = \frac{4 \cdot a_{k-1} \cdot a_k}{b_{k-1} \cdot g_{k-1}} \quad (7)$$

where

k = 2, 3,...n

But in addition we also need a_k and b_k:

$$a_k = \sin\left[\frac{(2k-1)\cdot\pi}{2n}\right] \quad (8)$$

where k = 1, 2, 3,...n

$$b_k = \gamma^2 + \sin^2\left(\frac{k\cdot\pi}{n}\right) \quad (9)$$

where k = 1, 2, 3,...n

We obtain the first coefficient for the "specimen deep pass" in the following way: If k = 1

$$a_1 = \sin\left[\frac{(2\cdot 1-1)\cdot\pi}{2\cdot 5}\right] = 0{,}309$$

This gives:

$$g_1 = \frac{2\cdot a_1}{\gamma} = \frac{2\cdot 0{,}309}{0{,}635} = 0{,}97322$$

For the next coefficient (k = 2), we need:

$$a_2 = \sin\left[\frac{(2\cdot 2-1)\cdot\pi}{2\cdot 5}\right] = 0{,}809$$

together with

$$b_1 = \gamma^2 + \sin^2\left[\frac{1\cdot\pi}{5}\right] = 0{,}748716$$

Thus as the second filter coefficient we obtain:

$$g_2 = \frac{4\cdot a_1 \cdot a_2}{b_1 \cdot g_1} = \frac{4\cdot 0{,}309\cdot 0{,}809}{0{,}748716\cdot 0{,}97322} = 1{,}37226$$

For the last coefficient, with

$$a_3 = \sin\left[\frac{(2\cdot 3-1)\cdot\pi}{2\cdot 5}\right] = 1{,}00$$

and

$$b_2 = (\gamma)^2 + \sin^2\left(\frac{2\cdot\pi}{5}\right) = (0{,}635)^2 + \sin\left(\frac{2\cdot\pi}{5}\right) = 1{,}3077$$

we finally obtain

$$g_1 = \frac{4 \cdot a_2 \cdot a_3}{b_2 \cdot g_2} = \frac{4 \cdot 0{,}809 \cdot 1{,}00}{1{,}3077 \cdot 1{,}37226} = 1{,}80324$$

8.3.2.2. Checking calculation results with Puff

The simplest method of checking the calculations would be to design a deep pass with any feasible, even limiting frequency, for example 100 MHz or 1 GHz. This is the easiest way to check the results on the screen.

But Puff also offers the option of incorporating the filter coefficients into the component table directly. Then you need only enter the desired impedance level and the limiting frequency in field F4, and the attenuation cycle for this case is plotted immediately. The components values actually required for the coils and capacitors can then be simply and rapidly determined using a pocket calculator.

As an example, let us do the calculations for an 100 MHz deep pass, with n = 5 and a maximum reflection factor of r = 10% in the "anti-coil format".

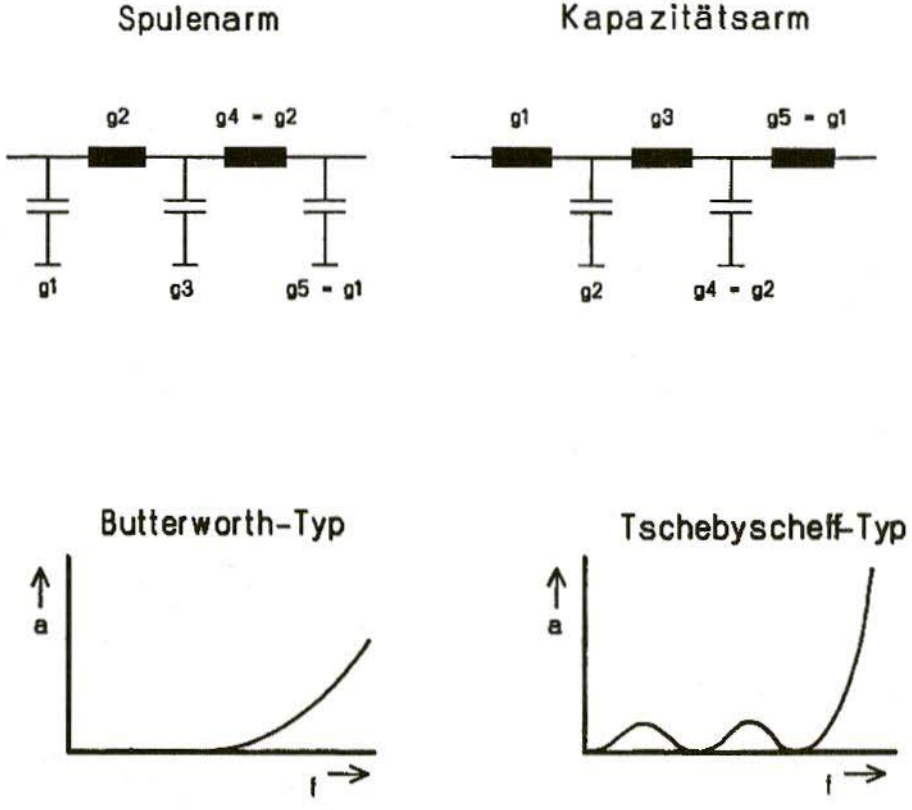
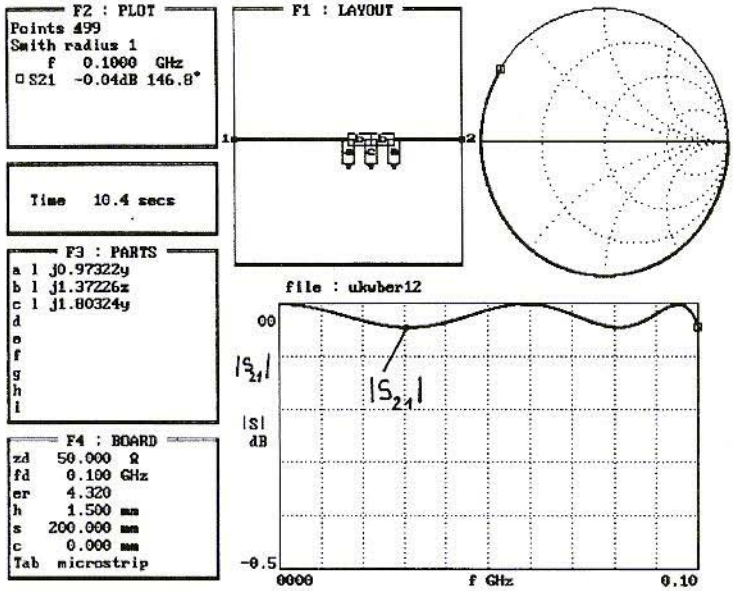


Fig.38: Standard Deep Pass in two Formats and Basic Transmission Loss Cycle
***Spulenarm* = Anti-coil, *Kapazitätsarm* = Anti-capacitanc", *"Typ* = Type**



of 100 MHz in the frequency response diagram, with a range of 0 to - 05dB in the vertical direction for S21, must obtain a print-out like the one in Fig.39.

If we now press the < page down> key and go right down to one of the attenuation maxima, we can read off a transmission loss of precisely 0.04dB in field F2, which was used in the calculations for this filter. If we now also plot the input reflection factor S11 for a value range switched from 0 to -50dB, for frequencies from 0 to 200 MHz, we will be amazed to see the pre-determined minimum reflection attenuation at 20dB, which is associated with a 10% reflection factor (Fig.40).

Fig.39: Simulation of a 100 MHz Deep Pass with n = 5 and r = 10%

To this end, the capacitors and coils are simply written into field F3 in standardised form - i.e. in the form of the coefficients g1, g2, g3 just calculated. But be careful. In this context, coils must be represented as "standardised reactive impedances". Capacitors, on the other hand, must be shown as "standardised susceptances". But a blind element is always accompanied by the letter "j". Standardisation on the impedance value is shown by the addition of a "z" after the component value, whereas standardisation on the susceptance is shown by the letter "y".

The F3 component list thus looks like this:

part a: lumped j0.97322y
part b: lumped j1.37226z
part c: lumped j1.80324y

Anyone pre-setting a limiting frequency

Determining the component value itself is relatively simple: Any nominated value from the F3 list is actually nothing more than the reactive impedance or susceptance of the component in question for the limiting frequency specified, based on the impedance value, here, of 50Ω. With this information, we can carry out the following calculations:

Anyone who wants to can now have fun

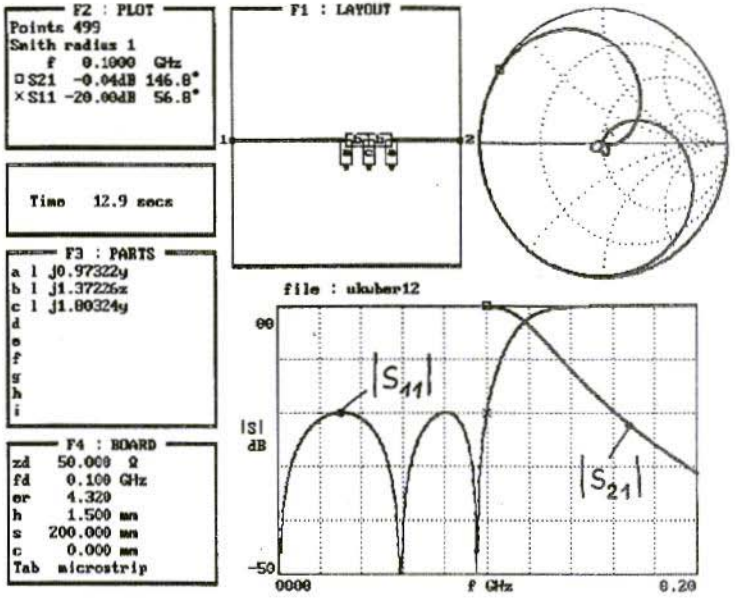


Fig.40: S11 and S22 for the 100-MHz Deep Pass, Range 0 to 200 MHz

$$Y_{C1} = \frac{0{,}97322}{50\ \Omega} \Rightarrow$$

$$\Rightarrow 2\pi \cdot f_0 \cdot C_1 = \frac{0{,}97322}{50\ \Omega} \Rightarrow$$

$$\Rightarrow C_1 = \frac{0{,}97322}{2\pi \cdot 100\ \text{MHz} \cdot 50\ \Omega} \Rightarrow$$

$$\Rightarrow C_1 = 30{,}97\ \text{pF}$$

$$X_{L2} = 1{,}377226 \cdot 50\ \Omega \Rightarrow$$

$$\Rightarrow 2\pi \cdot f_0 \cdot L_2 = 1{,}377226 \cdot 50\ \Omega \Rightarrow$$

$$\Rightarrow L_2 = \frac{1{,}377226 \cdot 50\ \Omega}{2\pi \cdot 100\ \text{MHz}} \Rightarrow$$

$$\Rightarrow L_2 = 109{,}2\ \text{nH}$$

$$Y_{C3} = \frac{1{,}80324}{50\ \Omega} \Rightarrow$$

$$\Rightarrow 2\pi \cdot 100\ \text{MHz} \cdot C_3 = \frac{1{,}80324}{50\ \Omega} \Rightarrow$$

$$\Rightarrow C_3 = \frac{1{,}80324}{2\pi \cdot 100\ \text{MHz} \cdot 50\ \Omega} \Rightarrow$$

$$\Rightarrow C_3 = 57{,}39\ \text{pF}$$

carrying out another Puff simulation with these components, and checking whether they really give the 100 MHz deep pass. Only SMD components may be used in the practical assembly of such a filter. Only they have such a high natural resonance (due to the tiny natural inductance of 1nH for the 0805 formats) that the resulting dips in the filter attenuation band no longer cause any interference. Of course, specific rules of play then have to be observed in designing the board, as well as in the screen housing construction, in order to obtain the attenuation losses predicted by the theory. Of course, we can now use Puff again to give a very precise prediction of the attenuation cycle going high up into the gigaHertz range, using the complete equivalent circuits of coils and capacitors, together with the stray capacitances of the solder flecks in the layout. *(To be continued)*

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Design and Realisation of Microwave Circuits Part-5 (cont)

8.3.2.3. Design of 1,700-MHz Stripline band pass, using the calculated values for the standard deep pass

If we start from a standard deep pass, with n = 5, and a reflection factor of 10%, we obtain the following filter coefficients:

g1 = 0.97322
g2 = 1.37226
g3 = 1.80324

The Stripline band pass filter here should be designed for an impedance level of 50Ω, with the lower limiting frequency, f_{bottom} = 1,680 MHz and the upper limiting frequency, f_{top}, = 1,780 MHz. This gives a band width of 100 MHz.

First step:

Determine the band pass filter coefficient, g_0, which now contains the limiting frequencies.

$$g_0 = \pi \cdot \frac{(f_{oben} - f_{unten})}{(f_{oben} + f_{unten})} \qquad (1)$$

This gives:

$$g_0 = \pi \cdot \frac{(100\ MHz)}{(3460\ MHz)} = 0{,}0907974$$

Second step:

Now you can determine the first Stripline pair; it has the same format as the last pair - i.e. no. 5.

First the "standardised impedance value":

$$\frac{Z_0}{k_{0,1}} = \frac{g_0}{\sqrt{g_0 \cdot g_1}} \qquad (2)$$

That gives:

$$\frac{Z_0}{k_{0,1}} = \frac{0{,}0907974}{\sqrt{0{,}09074 \cdot 0{,}97322}} = 0{,}30544$$

EVEN Resistance (= common mode resistance):

$$Z_{even} = 50\ \Omega \cdot \left[1 + \frac{Z_0}{k_{0,1}} + \left(\frac{Z_0}{k_{0,1}}\right)^2\right] \qquad (3)$$

$$Z_{even} = 50\ \Omega \cdot [1 + 0{,}30544 + (0{,}330544)^2] =$$
$$= 69{,}937\ \Omega$$

ODD Resistance (= push-pull resistance):

$$Z_{odd} = 50\ \Omega \cdot \left[1 - \frac{Z_0}{k_{0,1}} + \left(\frac{Z_0}{k_{0,1}}\right)^2\right] \quad (4)$$

$$Z_{odd} = 50\ \Omega \cdot [1 - 0{,}0544 + (0{,}30544)^2] =$$
$$= 39{,}3926\ \Omega$$

Third step:

We now come to the next line pair, no. 2 (it is identical to pair no. 4).

$$\frac{Z_0}{k_{1,2}} = \frac{g_0}{\sqrt{g_1 \cdot g_2}} =$$
$$= \frac{0{,}090744}{\sqrt{0{,}97322 \cdot 1{,}37226}} = 0{,}078568$$

This gives

$$Z_{even} = 50\ \Omega \cdot [1 + 0{,}078568 + 0{,}078568^2] =$$
$$= 54{,}237\ \Omega$$

and

$$Z_{odd} = 50\ \Omega \cdot [1 - 0{,}078568 + 0{,}078568^2] =$$
$$= 46{,}38\ \Omega$$

Fourth step:

Central line pair, no. 3.

$$\frac{Z_0}{k_{2,3}} = \frac{g_0}{\sqrt{g_2 \cdot g_3}} =$$
$$= \frac{0{,}0907974}{\sqrt{1{,}37226 \cdot 1{,}80324}} = 0{,}05772$$

Solution:

$$Z_{even} = 50\ \Omega \cdot [1 + 0{,}05772 + 0{,}05772^2] =$$
$$= 53{,}052\ \Omega$$

$$Z_{odd} = 50\ \Omega \cdot [1 - 0{,}05772 + 0{,}05722^2] =$$
$$= 47{,}28\ \Omega$$

Fifth step:

The three line pairs, with their even and odd resistances, are now entered into the F3 component list in Puff, together with the electrical length of 90° . With the complete circuit in field F1, a plot result as in Fig.41 must now be obtained.

We can see that a little "finishing up" is still needed to obtain uniform transmission attenuation waviness which will not be too high.

Obviously such great perfection was not yet possible at the time the original article was written - i.e. in the age of the slide rule and log tables - or else we would have come across the secret simplifications and / or short cuts of the authors of that time.

Naturally today's professional microwave CAD programs, such as Touchstone, SuperCompact, Harmonica, etc. can do everything much better, but of course they cost as much as a mid-range car!

Modern simulation and field analysis methods have taken design in this area to such a pitch of perfection that achieving the manufacturing precision levels for the track structures when the boards are engraved or prepared is like breaking the sound barrier. However, we shouldn't let that stop us, and should continue to experiment a little further with the calculated values. With a little effort and a few changes, we actually arrive at a cycle as in Fig.42.

Special attention should be paid here to obtaining a "ripple" as uniform as possible, for only in this way can you then obtain the lowest transmission attenuation when you plot the curve for

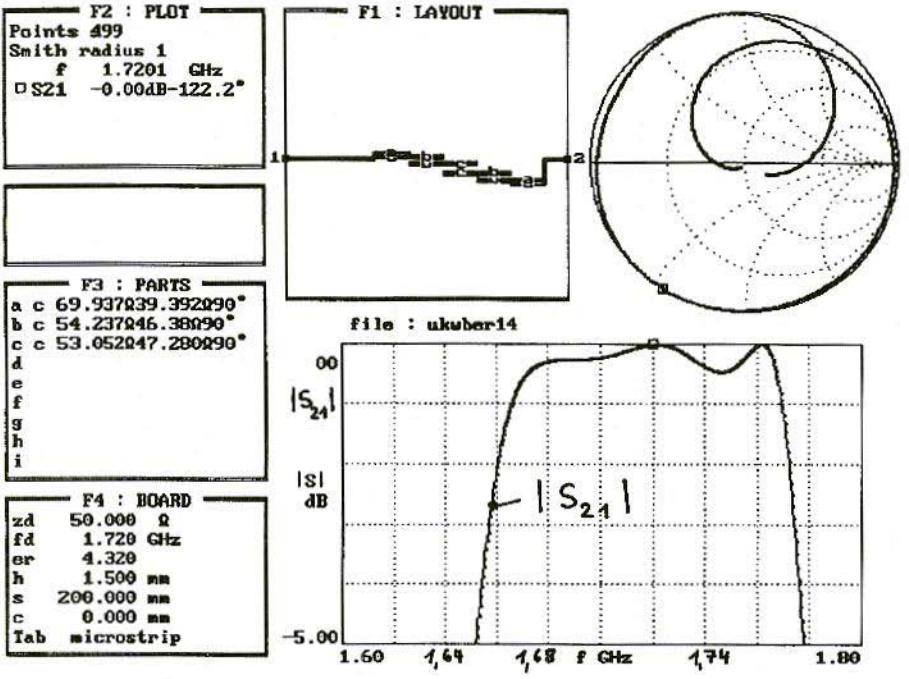

Fig.41: First Simulation of 1,700-MHz Band Pass with the Theoretically Determined Values

the filter with losses!

You have to think about this. If, as an experiment, we increase an even resistance anywhere in the table, the odd resistance should, as far as possible, be reduced by the same value, so that the impedance value of the line pair remains constant. Otherwise, you very quickly lose track.

In the next step, the unavoidable effects of harsh reality must be taken into account again.

So, we load the Puff file into the word processing program again in the usual way, check that all values, such as loss tangent, surface roughness, etc., are still correct, call Puff up again and match the line pair characteristics to reality.

To this end, we must put an exclamation mark in field F3 behind each "c" or "cl". Then the equals sign is keyed in on every line and as long as the electrical value of the component changes, until the resistances indicated in the dialogue field and the electrical lengths correspond to the ideal values of Fig.42.

When this has been done, we still need the mechanical dimensions of the tracks and their intervals, i.e. the "interaction gap" for the board design.

So we temporarily delete the exclamation mark in each component line and we enter the equals sign again. Back to the way things were!

We can now read off the length, width and intervals of the tracks in the

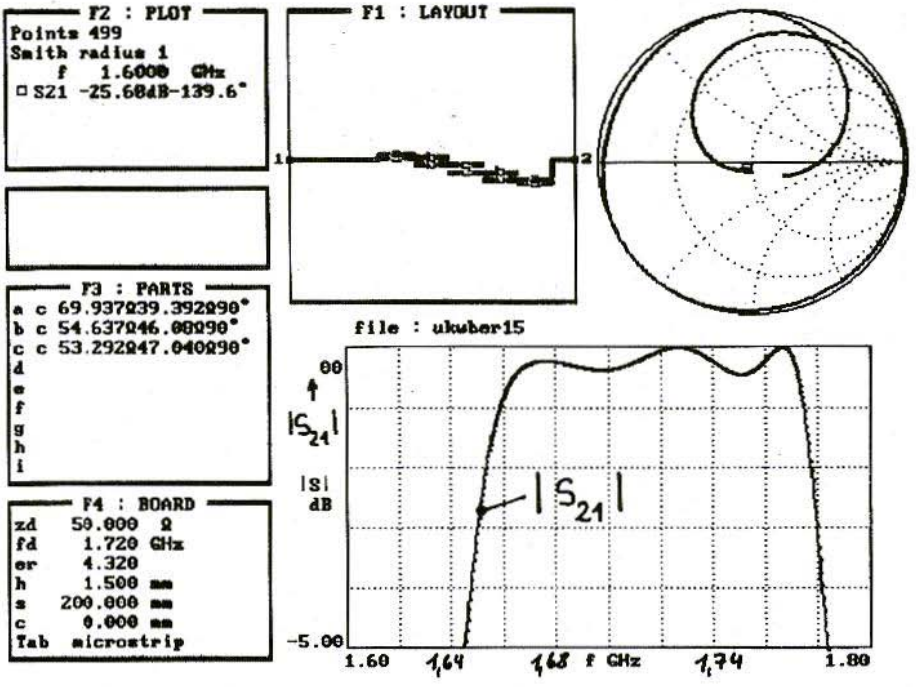

Fig.42: Improved Attenuation Cycle obtained by varying Line Data

dialogue field.

We can thus construct the following table:

(There is also the 50Ω supply line at the input and output, with a width of 2.88 mm..)

Line pair	Conductor width	Conductor length	Interaction gap
a	2.28	24.95	0.42
b	2.8	24.44	2.2
c	2.82	24.36	2.86

all dimensions in mm

The shortenings at the open ends and/or at the discontinuities between narrow and wide tracks still have to be determined. If we use the method from the last chapter to do this, we finally obtain the dimensions required for CAD design in accordance with Fig.43. Fig.44 shows the layout created with the Geddy CAD system, using a resolution of 0.02 mm.. The measurement results from the completed board, in a suitably milled aluminium housing with the external dimensions 35 x 135 mm. are entered in Fig.45, for the 1.5 to 2 GHz range, together with the simulation curve. Unfortunately, the measured transmission loss with 7.5 dB at 1,700 MHz is somewhat higher than the simulated result (approximately 6 dB). The pass range curve - as before, in the filter in the previous chapter - is again rather too narrow, whereas the resonator lengths and thus the limiting frequencies match up well.

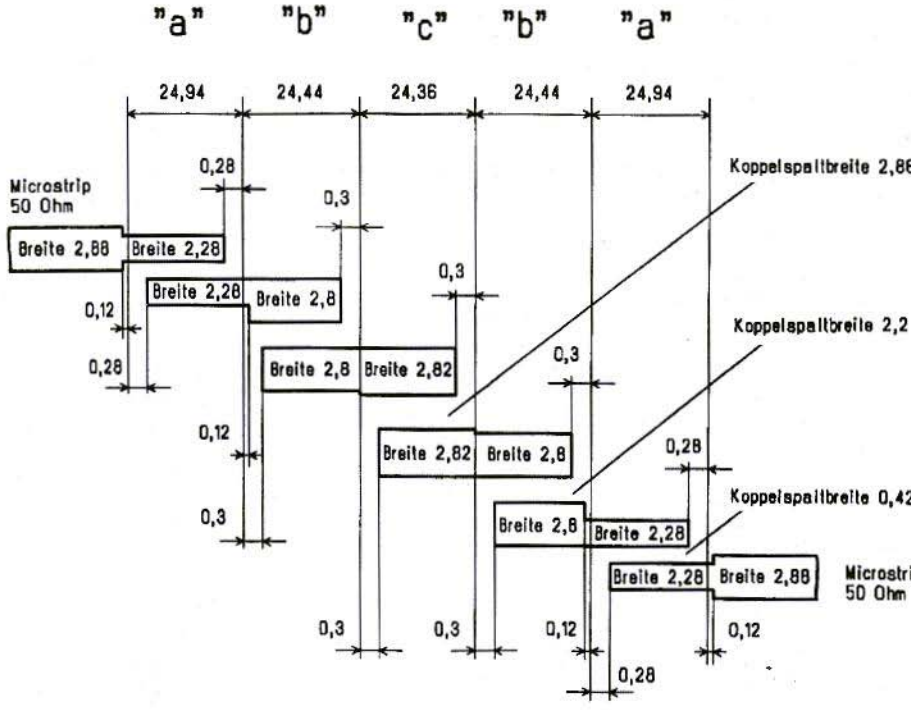

Fig.43: Dimensional Specification for Shortened Stripline
***Breite* = Width, *Koppelspaltbreite* = Interaction gap width**

Finally, Fig.46 shows the behaviour up to 10 GHz. Below 5 GHz, the correlation between theory and practice is almost perfect, whereas above this many previously unforeseen effects arise and reduce the filter attenuation.

In practice, we would also "frame" the band pass circuit with two appropriately dimensioned Stripline deep passes to guarantee, for example, a specific minimum attenuation value up to 10 GHz.

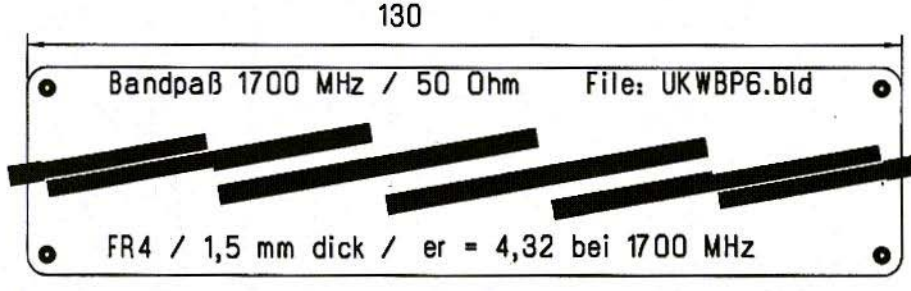

Fig.44: Completed Layout for 1,700-MHz Band Pass
***Bandpaß* = Band Pass", *Dick* = Thick, *Bei* = At**

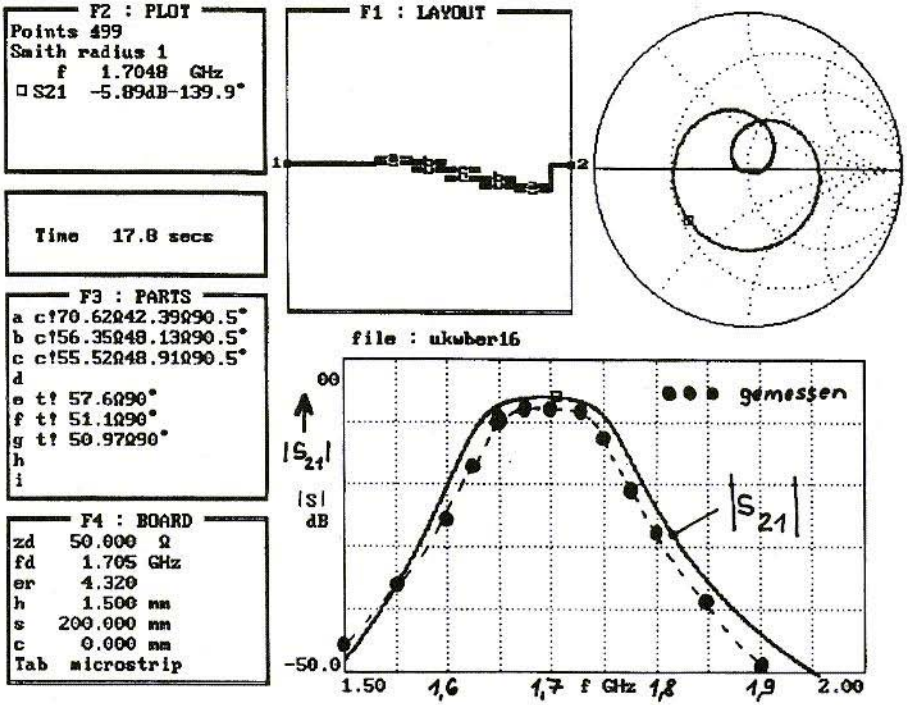

Fig.45: Simulation and Dimensional Results of Stripline Band Pass for f = 1.5 to 2 GHz and _S21_ = 0 to - 50 dB
Gemessen = Measured

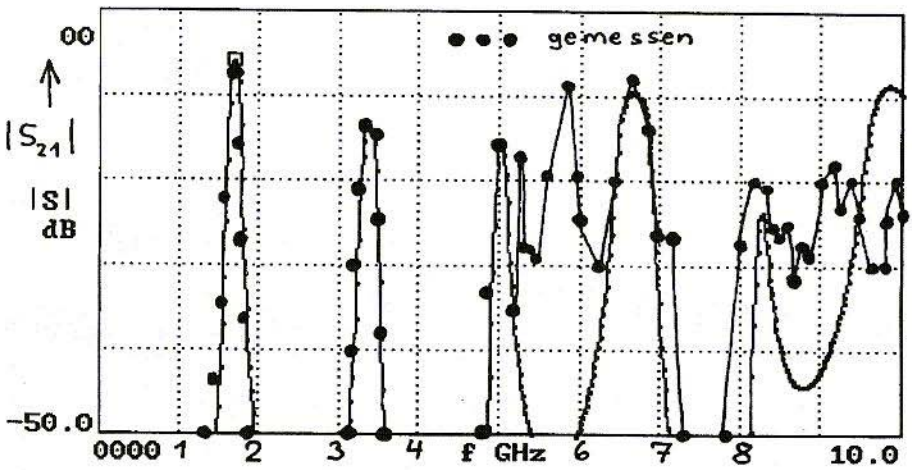

Fig.46: _S21_ in Range 0 to 10 GHz; *Gemessen* = Measured

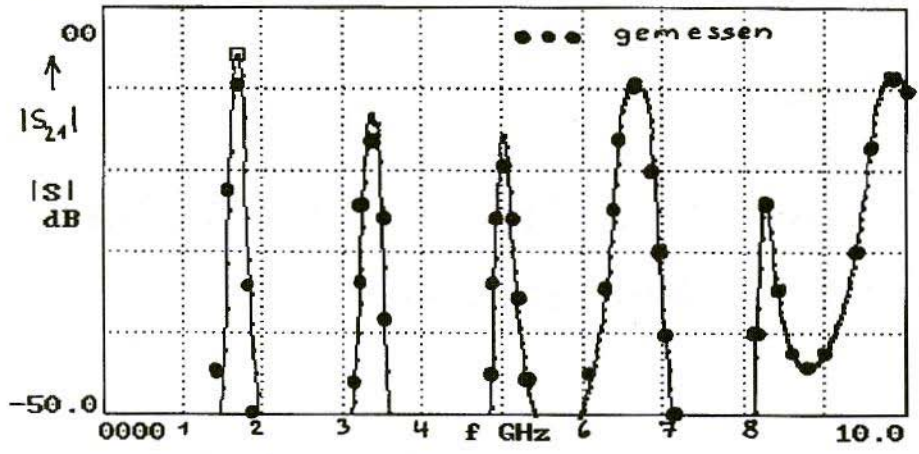

Fig.47: Cavity Resonances Suppressed by Conducting Foamed Material; S21 from 0 to 10 GHz of 1,700 MHz Band Pass Filter
***Gemessen* = Measured**

The beautifully milled aluminium housing would probably act as a wave guide and / or cavity resonator from 5 GHz. So the inner cavity over the board was lined, in that foamed material sections cut to size, using a conducting foamed material, were stuck to the inside of the cover. This was totally effective - just look at Fig.47! All the dips have disappeared, and the measured results up to 10 GHz are precisely on the simulation curve.

One little thing still needed to be corrected. Clearly the foamed material is a little too thick, and so is too near the Stripline. This shows in the fact that the attenuation calculated in the preceding transmission range at 1,700 MHz, and in the first scatter resonance (i.e. at 2 x 1,700 - 3,400 MHz) has increased by 1 to 2 dB. But this problem can certainly be solved by experiments with varying thicknesses of damping material and comparative measurements.

8.4. Literature on the subject

[1] S.B.Cohn: Parallel-coupled-Transmission-Line Resonator Filters
IRE Transactions on Microwave Theory and Techniques
MTT-6, April, 1958, pp. 223 - 231

[2] G.L.Matthaei: Design of Wide-Band (and Narrow Band) Bandpass Microwave Filters on the Insertion Loss Base
IRE Transactions of Microwave Theory Techniques
MTT-8, November, 1960, pp. 457 - 458

[3] Geschwinde & Krank: Striplines
C.F.Wintersche Verlagshandlung
Füssen, 1960, from p. 89

[4] Peter A.Rizzi: Microwave Engineering - Passive Circuits
Prentice Hall, New Jersey
ISBN 0-13-586702-9

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Design and Realisation of Microwave Circuits

Part-6

9. MICROSTRIP DEEP PASSES

9.1. Output Deep Pass after Mixer

Following the extensive band pass chapter, we now have to develop the two deep pass filters in the circuit

a) The intermediate frequency - i.e. the difference frequency between the input signal and the oscillator signal - should be screened out after the mixer. Thus the total frequency - the oscillator frequency including its second and third harmonics, as well as the total range of reception coming from the amplifier - must be suppressed. For this purpose, a deep pass is inserted here, with a limiting frequency of approximately 1 GHz.

b) To reduce the interference spectrum arising in the mixer, among other harmonics from the oscillator, there is a further deep pass between the oscillator and the mixer, with a limiting frequency of 1.7 GHz.

Both deep passes should be executed using stripline technology, with short sections of line with lengths of less than 10% of the wavelength as a replacement for discrete capacitors and coils as their extremely tiny values would lead to problems at higher frequencies with their natural resonances, or with precise calibration in series production. Stripline deep passes, however, have considerably greater dimensions to make up for this.

The following points apply here:

a) Wide, short sections of line with low impedance levels have high capacities, the inductances of which can still be neglected for short line lengths.

b) Short sections of line, as narrow as possible, with high impedance levels display a very small, even more insignificant capacity, and are therefore used as inductances here.

Of course, anyone using a board material with a dielectric constant which is too high - e.g. 10 - will have a nasty surprise here. Realising the inductances creates very big difficulties, since the

conductor widths can not be narrowed down indefinitely, and therefore the self-capacitance is always too high. It is best to stay below $\varepsilon_r = 5$.

It should also be pointed out that capacitors can be made effective in stripline technology with fewer problems than inductances. For this reason, we use the "low-coil" model of the already well-known Tschebyscheff standard deep pass, with n = 5, and a maximum reflection factor of 10%.

The filter coefficients were:

$$a_1 = 0.97322$$
$$a_2 = 1.37226$$
$$a_3 = 1.80324$$

By using these standardised values, we can, once again, implement any limiting frequency whatever very rapidly, and then easily specify the required component values.

If we initially start from the limiting frequency f_g = 1 GHz, we need only repeat the calculation from the previous chapter:

These component values should now be

$$Y_{C1} = \frac{0{,}97322}{50\,\Omega} \Rightarrow$$
$$\Rightarrow 2\pi \cdot f_g \cdot C_1 = \frac{0{,}97322}{50\,\Omega} \Rightarrow$$
$$\Rightarrow C_1 = \frac{0{,}97322}{2\pi \cdot 1000\,\text{MHz} \cdot 50\,\Omega} \Rightarrow$$
$$\Rightarrow C_1 = 3{,}097\,\text{pF}$$

$$X_{L2} = 1{,}377226 \cdot 50\,\Omega \Rightarrow$$
$$\Rightarrow 2\pi \cdot f_g \cdot L_2 = 1{,}377226 \cdot 50\,\Omega \Rightarrow$$
$$\Rightarrow L_2 = \frac{1{,}377226 \cdot 50\,\Omega}{2\pi \cdot 1000\,\text{MHz}} \Rightarrow$$
$$\Rightarrow L_2 = 10{,}92\,\text{nH}$$

$$Y_{C3} = \frac{1{,}80324}{50\Omega} \Rightarrow$$
$$\Rightarrow 2\pi \cdot 1000\,\text{MHz} \cdot C_3 = \frac{1{,}80324}{50\,\Omega} \Rightarrow$$
$$\Rightarrow C_3 = \frac{1{,}80324}{2\pi \cdot 1000\,\text{MHz} \cdot 50\,\Omega} \Rightarrow$$
$$\Rightarrow C_3 = 5{,}739\,\text{pF}$$

replaced by the sections of line. The relative length, l_c, of a short, wide line section, which is to act as a capacitance, is determined according to the following formula:

$$\frac{l_c}{\lambda} = f \cdot C \cdot Z_c$$

where "f" is the operating frequency and Zc the impedance level of the (wide) line used. We normally use values of between 10 and 20Ω for this impedance level.

A value as low as possible should be selected as this gives shorter line lengths, since the shorter the section of line is, the further the unavoidable "scatter resonances", with their attenuation interruptions in the filter attenuation band, slide away from the intended limiting frequency. True, this increases the conductor width, but its absolute limit is extended only by the width of the board or the screening housing.

Hence the experiment here with Z = 10Ω. For capacitor C1 this gives:

$$\frac{l_{C1}}{\lambda} = 1000\,\text{MHz} \cdot 3{,}097\,\text{pF} \cdot 10\,\Omega \Rightarrow$$
$$\Rightarrow \frac{l_{C1}}{\lambda} = 0{,}03097$$

To work with Puff, we need this electrical length in degrees. To do this, we merely need to multiply the value just obtained by 360, and thus obtain:

$$\mathbf{l_{c1} = 0.03097.360° = 11.15°}$$

The same procedure is applied to the capacitor C3:

$$\frac{l_{C3}}{\lambda} = 1000\,MHz \cdot 5{,}739\,pF \cdot 10\,\Omega \Rightarrow$$

$$\Rightarrow \frac{l_{C3}}{\lambda} = 0{,}05739$$

$$l_{C3} = 0{,}05739 \cdot 360° = 20{,}66°$$

Now we need "coils" again. In the narrow, high-impedance sections of line for the relative line lengths for producing an inductance, we have:

$$\frac{l_L}{\lambda} = \frac{f \cdot L}{Z_L}$$

Z_L represents the impedance level of the selected circuit for realising inductances. Here the conductor width should be selected to be as narrow as possible, and should go up to the practically realisable limit of 0.1 to 0.2mm.

Only in this way do we obtain the highest inductance values for the shortest circuit lengths, with the advantage of scatter resonances which are further away and higher filter attenuation.

Naturally, these extremely narrow paths increase the transmission loss in the pass band somewhat, but this is not important except for filters, which are used in transmitter high-level stages. Here there can certainly be problems - e.g. with heating up of the paths.

An impedance level of 135Ω is therefore selected here, which gives us:

$$\frac{l_{L2}}{\lambda} = \frac{1000\,MHz \cdot 10{,}92\,nH}{135\,\Omega} \Rightarrow$$

$$\Rightarrow \frac{l_{L2}}{\lambda} = 0{,}8088$$

In degrees:

$$I_{L2} = 0.08088.360° = 29.12°$$

The design procedure just discussed is summarised once again in Fig.48.

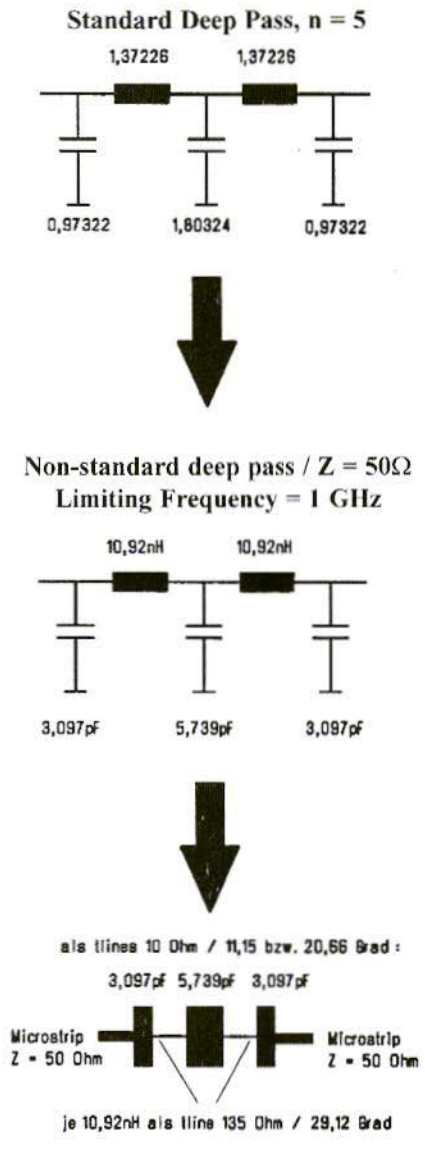

Fig.48: The Path from the Standard Deep Pass to the Microstrip Circuit
Als **= As,** ***Grad*** **= Degree,** ***Je*** **= Every**

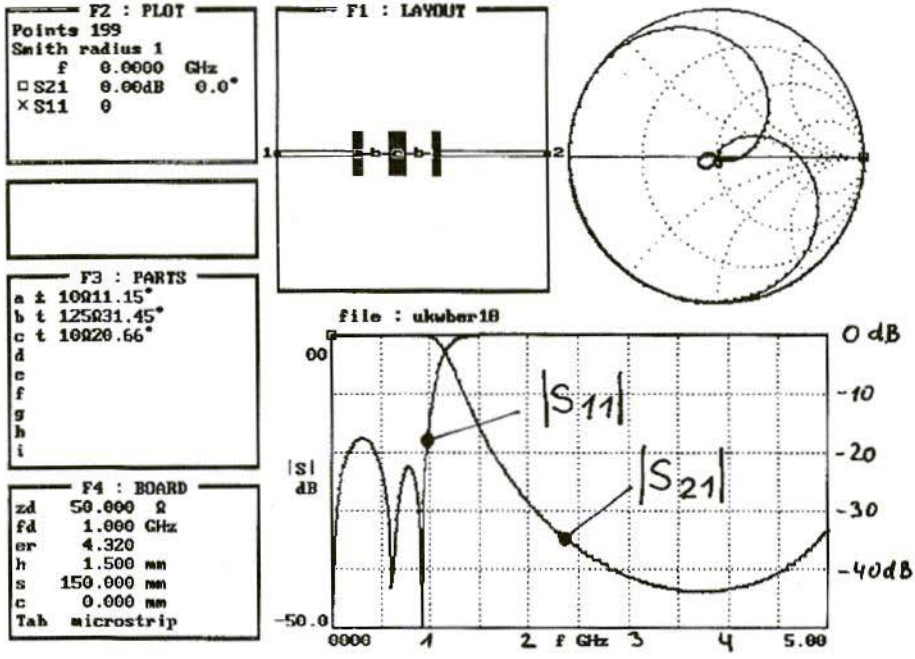

Fig.49: First Simulation of 1-GHz Deep Pass in Microstrip Technology (ideal circuit)

And now for the simulation using Puff. We enter the strip lines in field F3, prepare the layout in F1 and call up a plotting procedure. The inputs required are: frequency range 0 to 5 GHz, value range for |S11| and |S21| from 0 to -50dB. The result is shown in Fig.49, which naturally requires some explanation:

The values of the minimum input reflection attenuation |S11| are not equal, as between the poles, so that the procedure looks somewhat inelegant. This is simply due to the fact that we can unfortunately already observe the capacitive fraction in the "coils" - i.e. the wafer-thin lines - and we have to do something to counter this.

The solution is quite simple. We carefully reduce the length of the central line section "c" (which represents capacitor C3) in the component list F3 until the "unevennesses" have been eliminated. Check after each change by calling up a new plot.

If the minimum reflection attenuation is still rather too poor, we must also reduce to some extent the line sections "a" at the beginning and end of the filter, which form capacitors C1 and C5. A final correction at C3 then gives approximately the 20dB minimum reflection attenuation required.

If we continue, working with advanced modeling - i.e. with lossy sections of line, placing exclamation marks behind

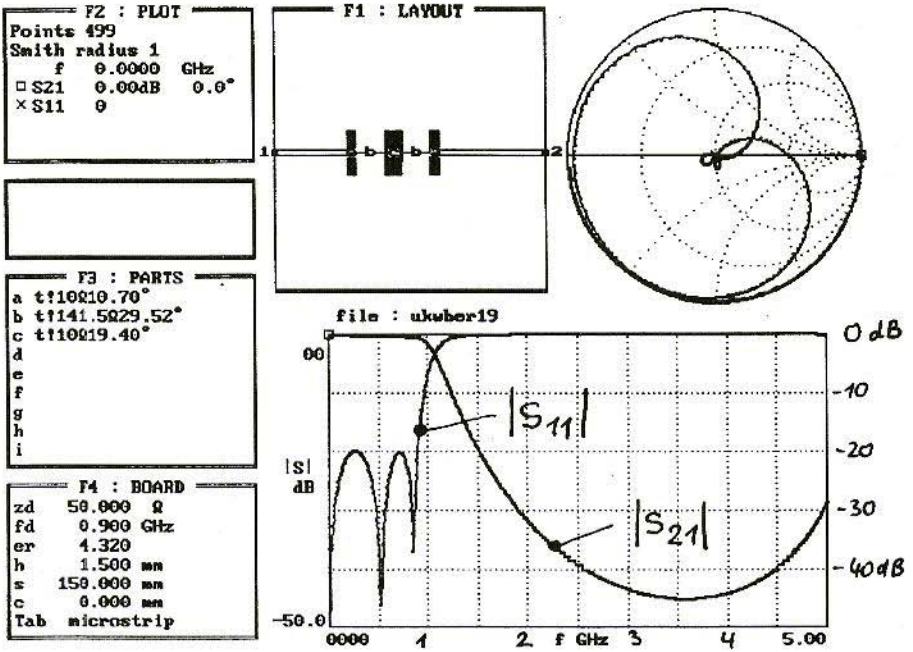

Fig.50: GHz Deep Pass with Losses following fine correction

the "tline" values again, and correcting the earth until the ideal values are displayed again, we obtain Fig.50.

Here the limiting frequency has just been decreased a little further to obtain at least 20dB filter band attenuation at the oscillator frequency of 1,600 MHz. We have also made sure that the third harmonic of the oscillator frequency (4,800 MHz) was also sufficiently suppressed. This is ensured if app. 20dB of attenuation is present there as well.

Now it's the well-known procedure again: the open-end shortening of the wide line sections is determined again, the narrow lines are lengthened by these abbreviations, feed lines are connected with Z = 50Ω, and the layout is designed. The diagram in the Puff manual (page 37) already mentioned gives us a shortening of 0.47 x board thickness for 10Ω and ε_r = 4.32. For a thickness of 1.5mm, this gives, fairly accurately, 0.7mm.

The necessary CAD table is then as described by Table 1.

The 50Ω feed requires an additional width of 2.86mm.

Note:

Different open-end corrections are necessary for line section "a" (for C1 and C5), for it is connected with the "inductance" on one side only. The tiny width of 0.22mm at "b" need not be taken into account, as against 23.68mm, so on this side the complete shortening applies of 0.47 x h = 0.47 x 1.5mm = 0.7mm.

But the width of the 50Ω line connected

Section	Z	CAD-Width	PUFF-Length	open-end Correction	CAD-Length
a	10Ω	**23.68 mm**	4.98 mm	see special note	**see special note**
b	125Ω	**0.22 mm**	16.16 mm	+2 x 0.7 mm = +1.4 mm	**17.56 mm**
c	10Ω	**23.68 mm**	9.02 mm	-2 x 0.47 x h = -2 x 0.7 mm	**7.62 mm**

Table 1

to the other side is 2.86mm. So here we need correct only by:

$$\Delta l = 0{,}7\,mm \cdot (1 - \frac{W1}{W2}) =$$

$$= 0{,}7\,mm \cdot (1 - \frac{2{,}86}{23{,}68}) = 0{,}62\,mm$$

Thus, for both the first and last sections of line, we obtain a length of 4.98mm - 0.7mm - 0.62mm = 3.66mm, with a width of 23.68mm. Fig.51 shows the layout thus created.

As a comparison, we also investigated the influence of the foamed material glued to the underside of the housing cover to suppress the wave guide effects and cavity resonances.

Fig.52 shows the transmission loss simulation and measurement result for the frequency range from 0 to 1 GHz.

To sum up, we can say the following:

a) The limiting frequency measured, which was rather lower than the theoretical value, and the rather high degree of ripple for the attenuation in the transmission band can be traced back, for the most part, to the fact that the dielectric constants in the FR4 circuit board material are dependent on the frequency, which means the component values obtained are not precisely what they should be.

This made an immediate impression. At low frequencies, ε_r was still 4.8, and had fallen to 4.3 at 1.7 GHz. If we call Puff up again and check through which

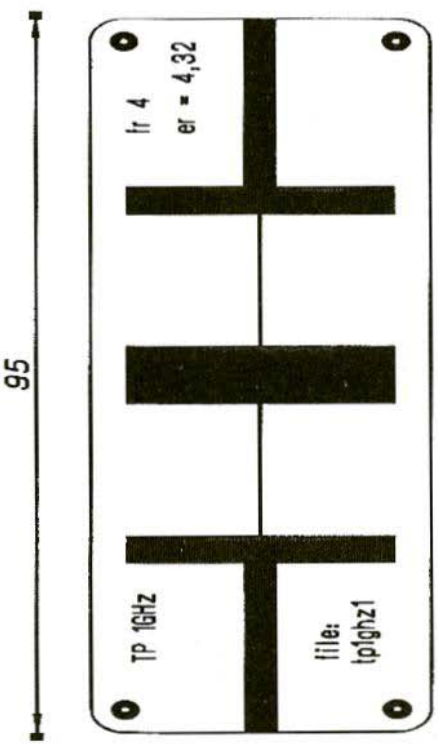

Fig.51: Layout of 1 GHz Deep Pass

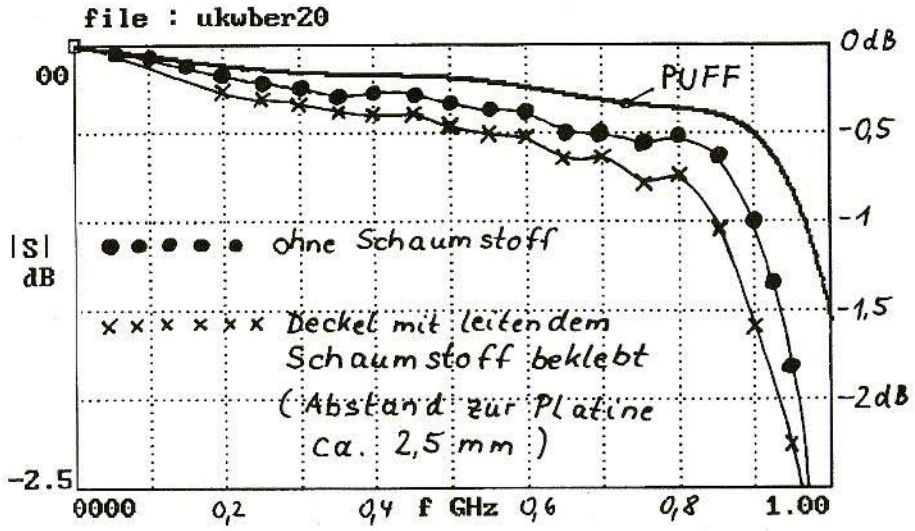

Fig.52: Deep Pass Transmission Attenuation: Range 0...1 GHz; Additional Curves: Influence of Attenuation Material
Ohne Schaumstoff = Without foamed material, *Deckel mit leitendem Schaumstoff beklebt* = Cover glued on with conductive foamed material, *Abstand zur Platine* = Distance from board, *Ca.* = Approximately

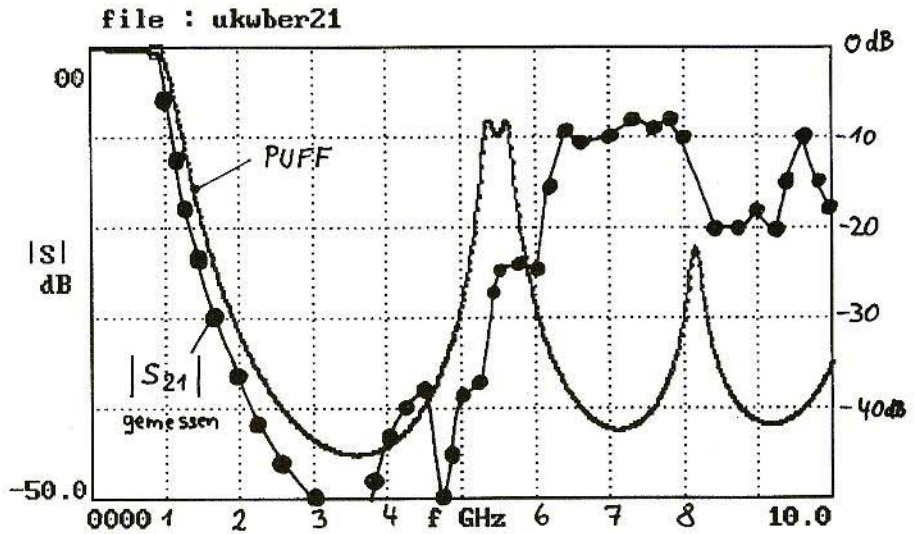

Fig.53: Path of Filter Attenuation of Microstrip Deep Pass, Range 0 to 10 GH
Gemessen = Measured

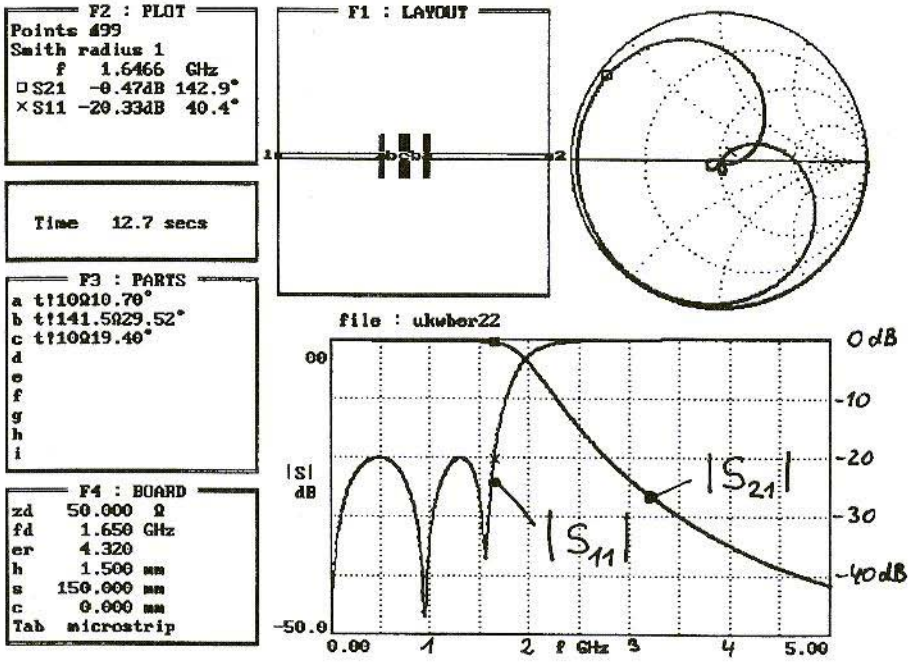

Fig.54: Simulation of 1.7 GHz Deep Pass with Losses following Displacement of Design Frequency

dielectric constants are well matched to the curve path measured, then we come to 4.7 at 900 MHz.

Those who want to work it out for themselves should do as follows. In component list F3, replace all length specifications in degrees by specifications in mm, which have already been entered in the tables above for CAD work. Then change the value for er in field F4, without changing the reference frequency, and continue to plot the new values until the simulated path and the measured result coincide.

b) The conducting foamed material on the underside of the cover certainly attenuates the cavity resonances. Unfortunately, it also increases the transmission loss, which is approximately 0.3 - 0.4dB at 900 MHz, with a distance of app. 3mm between the foamed material and the board surface. The smaller the distance between the board and the attenuation material and the higher the frequency, the greater the additional attenuation.

Finally, repeat the measurements for $|S21|$ in the frequency range from 0 to 10 GHz (Fig.53). Here again it can clearly be recognised that the component measurements - probably because er has been given as 4.3 instead of 4.7 - are too high, and so the increase in the attenuation from the transmission limit begins too early.

Section	Z	CAD-Width	PUFF-Length	open-end Correction	CAD-Length
a	10Ω	**23.68 mm**	2.70 mm	-(0.7 + 0.62) = -1.32 mm	**1.38 mm**
b	125Ω	**0.22 mm**	8.82 mm	+2 x 0.7mm = +1.4 mm	**10.22 mm**
c	10Ω	**23.68 mm**	4.92 mm	-2 x 0.47 x h	**3.52 mm**

Table 2

9.2. Oscillator Deep Pass

This assembly follows the oscillator circuit, and should still allow the oscillator frequency of 1.6 GHz to be well attenuated. But it should prevent the harmonics (3.2 GHz, 4.8 GHz, etc.) from getting into the subsequent mixer circuit and leading to undesirable mixed products there.

To do this, we call up the Puff file from the previous circuit, alter the reference frequency in field F4 to 1.65 GHz, and obtain Fig.54.

Using component list F3, after briefly deleting the exclamation mark, and entering the equals sign, we return to the electrical dimensions of the stripline sections. In addition, at this frequency - about 1.7 GHz - we can reckon on er = 4.32 again in fact.

If we then carry through the open-end correction again, with the values 0.7mm and / or 0.62mm, we already have the data to design the board (Table 2).

The 50Ω line again requires a width of 2.86mm.

If we examine the list carefully, we see that the CAD outputs become so exotically brief - due to the low impedance levels of 10Ω selected and a resulting width of 23.68mm for the "capacitors" - that the open-end shortening is already of the order of the residual length.

This is naturally a very tricky area, due to the dispersion of the board material data, and because of the risk of underetching the track edges while preparing the board.

Fig.55 shows the board design, which is already displaying a rather unusual look,

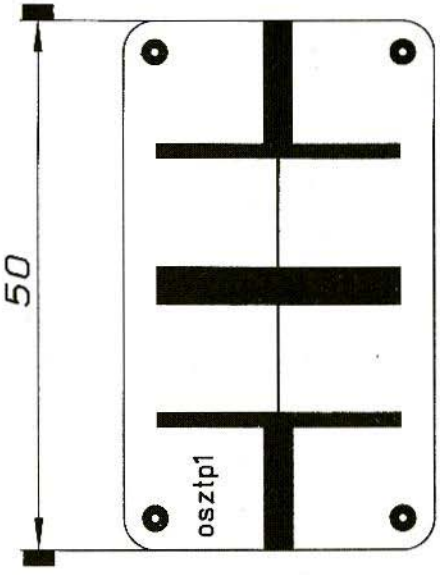

Fig.55: Layout of 1.7 GHz Microstrip Deep Pass, First Version

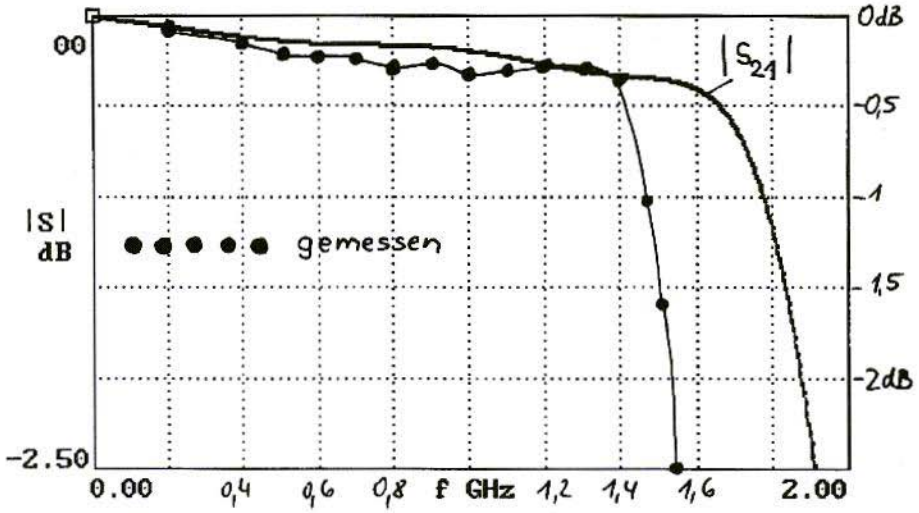

Fig.56: Transmission Loss |S21| in Frequency Range 0 to 2 GHz for First Version of Deep Pass
***Gemessen* = Measured**

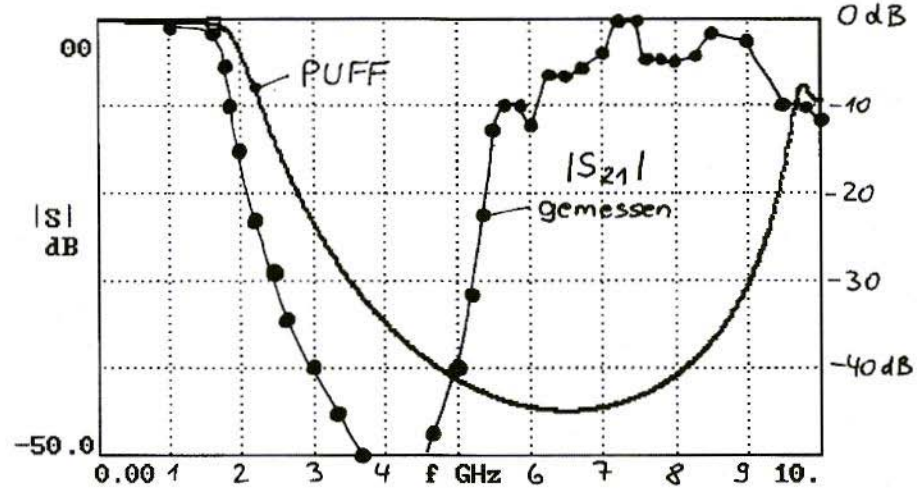

Fig.57: |S21| for Range 0 to 10 GHz in 1.7 GHz Deep Pass, First Version
***Gemessen* = Measured**

for the whole thing almost looks like a Yagi antenna.

The measurement results look just as unpleasant, for whilst the transmission loss measured corresponds very precisely to the calculated value (see Fig.56), the transition from the transmission band to the filter attenuation band

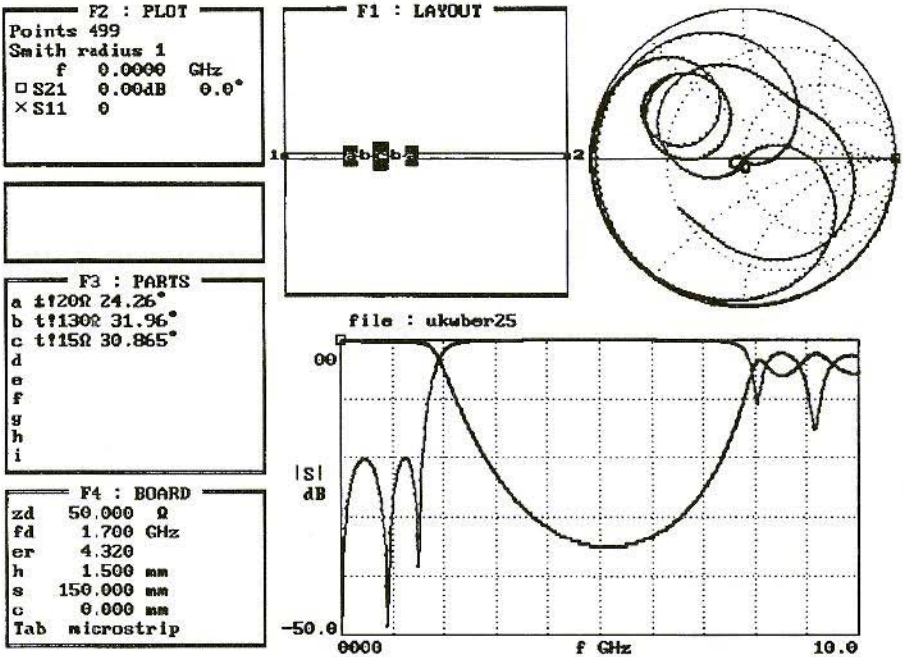

Fig.58: Puff Simulation of 1.7 GHz Microstrip Deep Pass, Second Version

sets in markedly too early. Anyone looking more closely at the diagram can see that there is a deviation of about 250 to 300 MHz, as against the theory - i.e. about 15 to 20% - decidedly too much!

Measurement up to 10 GHz (Fig.57) confirms even more clearly that the transition is far too early, to which are also added, from app. 5 GHz, the usual effects, such as cavity resonance, wave guide behaviour, re-radiation from the track edges, etc., which can not be detected by the simple Puff simulation model being used here.

But the markedly steeper increase in the filter band attenuation from the transmission limit onwards must be evaluated as genuinely positive.

Unfortunately, the attenuation collapses completely from 5 GHz - there is probably not much filter attenuation to speak of here.

One way out which offers itself is an increase in the impedance levels in the "capacitor striplines", for this reduces the track widths and thus the lines must be extended.

The impedance levels at C1 and C2 are experimentally doubled from 10 to 20Ω, while the impedance level at the middle capacitor is increased only from 10 to 15Ω. Its capacity is in fact twice as high as that of C1 and C5, and the conductor length should be kept as short as possible, to keep the "scatter resonances" within bounds.

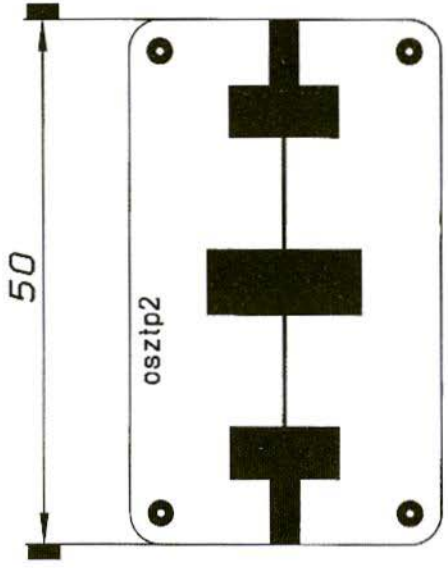

Fig.59: Layout of 1.7 GHz Deep Pass, Second Version

Even the thin lines for the "inductances" are widened, in that the impedance level is lowered to 125Ω. This makes it easier to handle the problems arising when boards are produced with such long, wafer-thin lines.

The filter calculation and layout, including the fine correction, is once again carried out exactly as in the method described above. The results and / or the component values for the "ideal" deep pass are shown in Fig.58. As expected, the attenuation rapidly falls back above 5 GHz, because the scatter resonances are favoured by the greater conductor lengths. In contrast, the layout shown in Fig.59 works out to be somewhat more balanced. The diagrams in Figs. 60 and 61 include the losses, and the values measured on the board are also entered.

The result now looks a little better, for the deviations from the simulation lie at around 10%, if we deduct the rather higher basic attenuation due to the mounting of the board in the housing, including the junction between each SMA jack and the board.

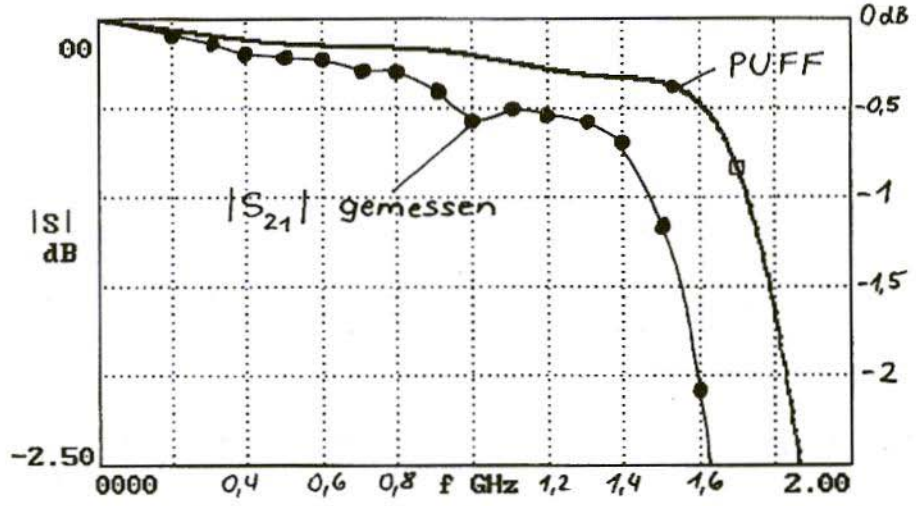

Fig.60: Measured Path of Transmission Loss in Range 0 to 2 GHz of 1.7 GHz Deep Pass, Second Version
***Gemessen* = Measured**

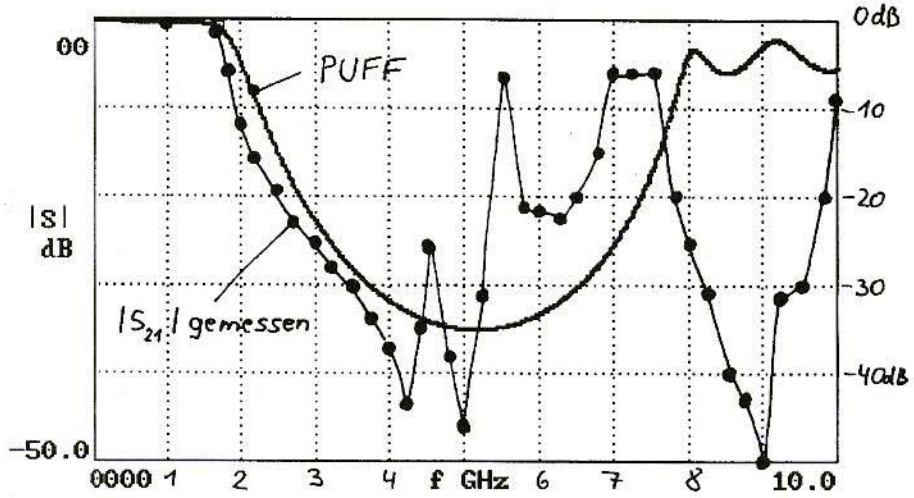

Fig.61: |S21| in Range 0 to 10 GHz of 1.7 GHz Deep Pass, Second Version
***Gemessen* = Measured**

Here too, we can make a quick end of matters, using the following method:

We estimate how many MHz too early the rise to the filter attenuation band begins. In the deep pass board just examined, this is approximately 170 MHz.

We simply increase the limit frequency in field F4 by this amount - so we enter 1,820 MHz there.

Then we use the component list F3 and the equals sign to obtain the new associated mechanical dimensions of the three striplines a to c. After open-end shortening of the wide lines, plus the extension of the "inductances", a new corrected layout is prepared and a board produced.

The measured results obtained in this way are shown in Figs. 62 and 63. There is now a very serviceable match, but of course, once again, only up to about 5 GHz.

9.3. Summary of Experiences and Design Rules for Micro-Strip Deep Passes

1. Start with the standard deep pass, re-dimension it to the desired operational values and calculate the values for the required coils or capacitors.

2. Capacities are realised through very wide striplines (initial value: Z app. 20Ω), inductances by means of very narrow lines (guideline: Z to exceed 100Ω by as much as possible).

3. The line lengths here should not exceed 10% of the wavelengths (corresponding to an electrical length of 36°).

4. In the first Puff simulation, we almost always observe a marked deviation from the ideal path for the S parameter, which is due to the capacitive fraction of the "inductance lines". By carefully reducing the line lengths on the "capacities" under continuous "Puff monitoring", this defect can be eliminated.

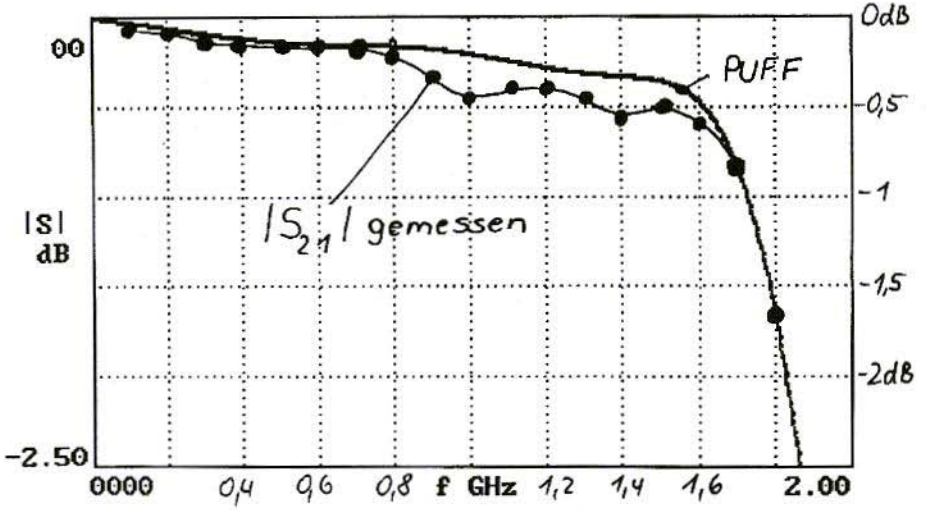

Fig.62: |S21| for 1.7 GHz Deep Pass, Third Version, but with Design Frequency increased by 170 MHz
Gemessen = Measured

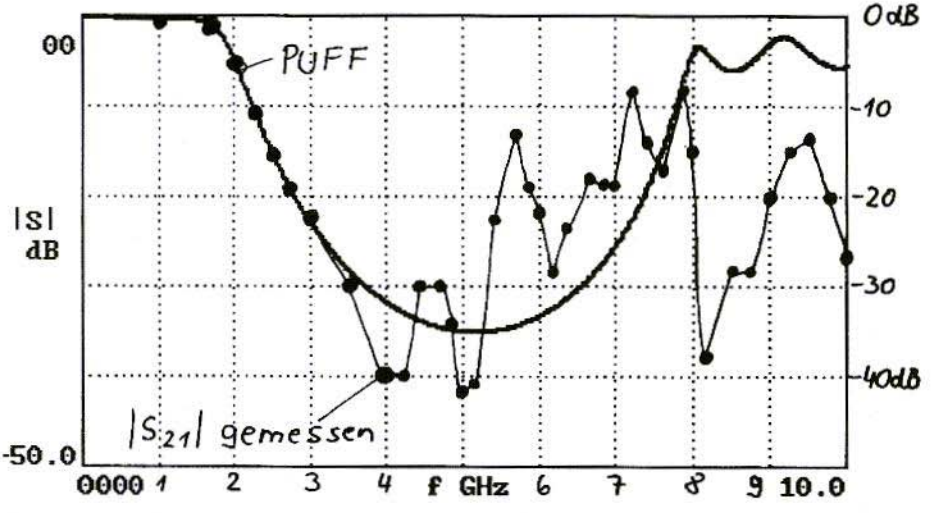

Fig.63: As Fig.62, Range from 0 to 10 GHz
Gemessen = Measured

5. We should try and make the line lengths for the "capacitors" as short as possible, for only in this way can the unavoidable interruptions in the filter attenuation band (and thus the scatter resonances) be pushed very high up.

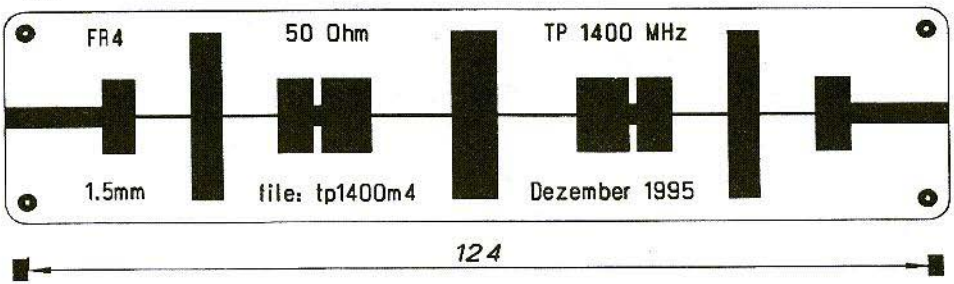

Fig.64: Expanded 23cm Deep Pass with Improved Filter Attenuation
***Dezember* = December**

However, it is not just the board or housing dimensions which set the limits for expanding the line widths of microstrips. As the line lengths increase and the widths grow, structures actually arise which can no longer be correctly represented by simulations using Puff. The difference between theory and practice rises to more than 20%. Moreover, in certain circumstances, the attenuation collapses to zero in the filter attenuation band.

So from experience we can recommend that in the "capacitor lines" the ratio between the track width and the final track length on the board should be selected to be under 8 to 10.

6. When the completed filter is measured, we always detect a displacement of the actual limiting frequency towards lower values. The frequency deviation should then be determined, and the simulation should be repeated, with a design frequency in field F4, increased by the amount of the error.

When the new dimensions have been determined, carry out the correct procedure for an open end correction and re-jig the layout and board, and you will usually be on target.

7. Should there be interference from the interruptions in the filter attenuation band (= the scatter resonances), we can combine the desired deep pass, individually, on the left-hand and right-hand sides, with a further additional deep pass. These additional filters have limiting frequencies which are so high that their filter band attenuation comes in useful where the desired pass is already displaying its interruptions.

Fig.64, a deep pass for the 23cm band, shows how such a solution might look in practice.

8. The following rules generally apply in the selection of boards and materials in the micro-wave band:

– The higher the design and operational frequencies are, the thinner the board material must be selected to be, so as not to favour the re-radiating of energy at the track edges.

– Material with a thickness exceeding 1mm is normally used only in the range under 2 GHz. Even FR4 material should be used only up to this limit. Up to this frequency, it also makes sense to select a copper coating with a thickness of 35µm. For higher frequencies, the skin

effect means the copper layer can be selected to be thinner without any reservations.

– Boards made from pure Teflon material are appallingly impractical to deal with. The advantages of the extremely low losses up to and beyond 10 GHz have to be paid for by inadequate mechanical properties. Not only does it take a lot of skill to drill the holes (because the drill does not meet any kind of resistance). There is also the fact that the board bends even under its own weight.

– Mounting within the housing can be done only by soldering to the continuous-earthed underside, or by using very expensive epoxy conducting adhesive, which contains a lot of silver. There are just as many problems in connection with the through-plating of the earth connections, since first you have to do a little etching with hydrofluoric acid.

– The later developments (e.g. Teflon or ceramic powder with glass cloth) are rather more good-natured, and mechanically more stable. There are a few recent developments, e.g. from Rogers, the "RT-Duroid inventors", which point to really tremendous progress.

We can now use strongly cross-linked hydrocarbons as mechanical stabilisers in the carrier material and, with the same or better quality, as against RT-Duroid, we obtain such high strength that such boards can be sawn, drilled, through-plated and screwed exactly like FR4 material (i.e. epoxy glass cloth).

To be continued

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Design and Realisation of Microwave Circuits

Part-7

10. MICROWAVE MIXER STAGES

10.1. Principles

The word "mixing" is a rather unfortunate way of describing the function of this stage. All mixer stages are actually amplitude modulators.

A "tone" signal, which we shall henceforth refer to as the high-frequency signal (RF signal), is amplitude-modulated through a "carrier signal", which is usually generated by a carrier oscillator hence the description "LO signal" in the literature. In the ideal mixer, thanks to this amplitude modulation, only the total and difference frequencies are generated, whilst all other frequencies are completely suppressed (Fig.65).

Both these signals are of interest here, for we can mix upwards (filtering out the total frequency) or mix downwards to form a lower intermediate frequency, as usually occurs in a superhet receiver.

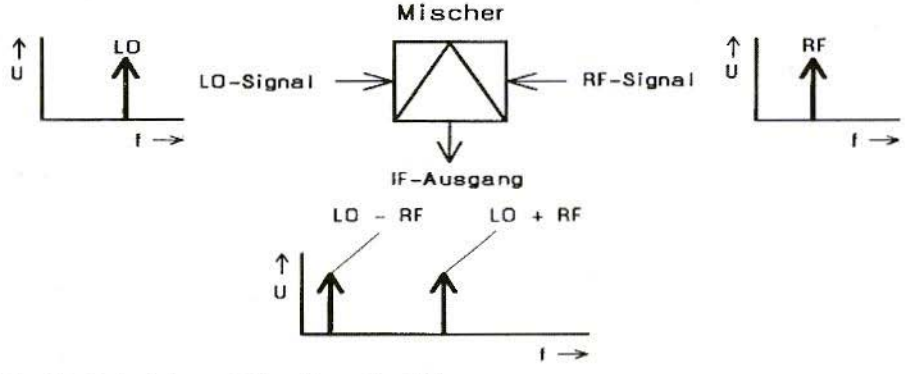


Fig.65: Principle and Function of a Mixer
***Mischer* = Mixer, *IF-Ausgang* = IF output**

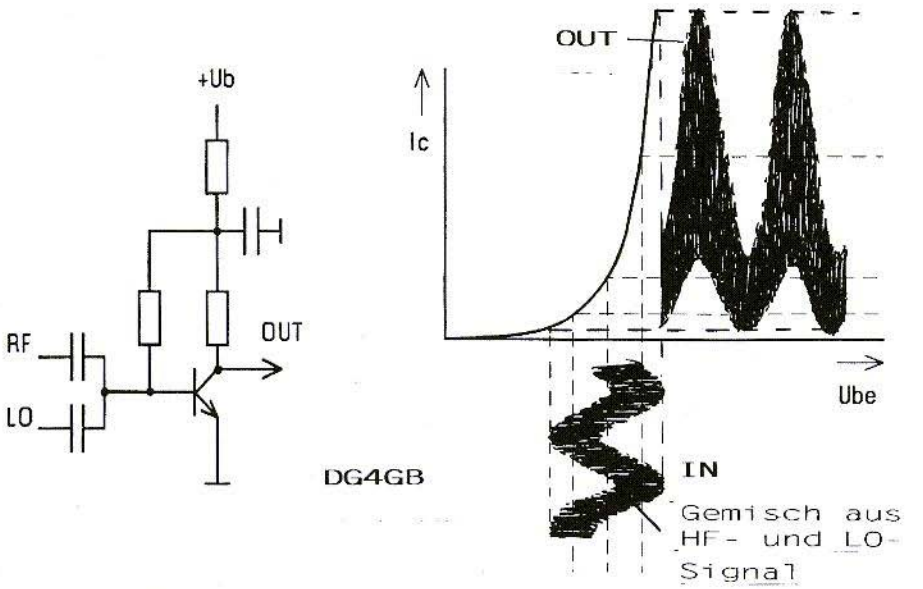


Fig.66: Additive Mixing on a Curved Characteristic
***Gemisch aus HF- und LO-Signal* = Mixture of HF and LO signals**

10.2. Additive Mixer Stages

They probably represent the oldest form of mixer, for all that is needed is any component with a curved characteristic or, to put it more precisely, with a non-linear relationship between the current and the voltage or between the input signal and the output signal.

It makes no initial difference whether this is a diode, a transistor, a field effect transistor, a voltage-dependent resistor or a filament bulb.

If both voltages are simultaneously applied to the component selected as mixer the LO signal usually has a higher amplitude this LO signal now continuously pushes the operating point to and fro along the characteristic. So the characteristic gradient for the smaller high-frequency signal is permanently changing and the amplitude modulation required for the high-frequency signal fraction is already present at the output!

Of course, no amplification can be expected from the simple mixing circuits with diodes. However, even today this method is still used for mixing in many Astra LNC's with Schottky diodes. Diode mixers in the dwarf wavelength range will be with us for some time yet.

If, on the other hand, the basis emitter segment of a bipolar transistor or the gate source segment of an FET is used as the modulator element, things are already looking much better with regard to amplification.

Since 1930, millions and millions of superhet receivers have demonstrated that this technology works well (see Fig.66).

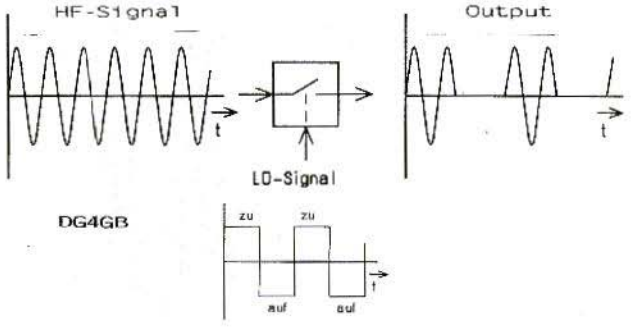


Fig.67: Simple Multiplicative Mixer

However, if we now look at the precise gradient of the output signal, we can immediately recognise the main disadvantage of this very simple principle the curved control characteristic does not yield textbook AM but something which is decidedly distorted. So in addition to the total frequency and difference frequency required, this signal gradient also contains both the input signals fed in (LO and HF), together with innumerable harmonics and their combination frequencies.

The only way to obtain acceptable inter-modulation ratios is consequently to work with low signal levels and good filtration.

A pure diode mixer stage unfortunately also has two additional secondary effects: Firstly, the attenuation already referred to must be compensated by additional amplification. Secondly, attenuation always leads to an increase in the noise factor of this assembly by the amount of attenuation.

Thus we need a very good, low-noise pre-amplifier with still higher amplification and a very conscientious check calculation to determine the total noise factor of the system.

10.3. Multiplicative Mixer

It would be better to identify this model through its most important characteristic. We are actually talking about the relay mixer group. In the most simple type of multiplicative mixer, the simple modulator, the high-frequency signal is "chopped up" to fit the rhythm of the LO signal (Fig.67). A square-wave modulation is generated in this way, in which the undesirable signal fractions and interference products in the output signal are reduced by 20 to 30dB, as against additive mixing.

Of course, we now need a considerably higher LO voltage, which must usually have a peak-to-peak value of at least 1 to 1.5 V, which is equivalent to a minimum level of + 7dBm. A genuine square-wave voltage of this amplitude, with flanks as steep as possible, is even better than the usual sinusoidal form. The electronic switch must actually be opened in a flash to fit the rhythm of the carrier frequency and must be closed again just as fast. Only in this way can we arrive in the initial control characteristic segment with the strongest curvature, which also produces the most distortions, as quickly as possible. This naturally has a more favourable effect overall on the distortions arising.

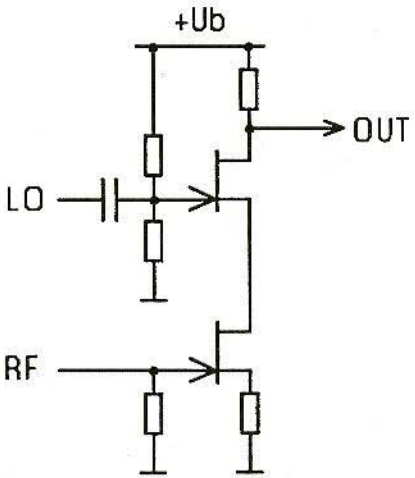


Fig.68: Modern Multiplicative GaAsFET Mixer

But there is one disadvantage we should not hide. Neither the HF signal nor the LO signal is suppressed in this circuit. They emerge at the output and must be eliminated with filters. By using active components such as transistors or FET's, mixing amplification can be obtained with this circuit as well.

A simplified version of the inner workings of a modern, integrated mixer module constructed according to this operating principle can be seen in Fig.68. It is easy to recognise the two GaAsFETs, the lower one of which operates as an amplifier for the high-frequency signal. The other one, by contrast, acts as an "on/off switch, activated by the LO signal".

The circuit will certainly look familiar to one or two people. It is a "well-known cascade connection".

This circuit is at present replacing the Astra LNCs at breakneck speed i.e. all kinds of passive diode mixers at 12 GHz, since they provide considerable amplification. At the same time, the FET acting as "chopper up" serves as the conversion oscillator. This means additional components can be saved on and the costs can be reduced.

However, anyone wanting to come as close as possible to the ideal mixer must resort to the balanced mixer. This is well known as the ring modulator or double balanced mixer, with 2 transmitters and 4 diodes. It not only chops signals up, but switches poles through 180° as well (Fig.69). This balanced principle not only reduces the harmonic content further but also reduces the suppression of the two input signals. If we now also take it into account that the LO signal acting as the turn-on voltage should, as far as possible, be 20dB above the high-frequency level, there is not much left to be desired in a mixer.

Naturally, the double balanced mixer can also be created using active components. A distinction should certainly be made between two methods:

a) The diodes in the circuit are replaced, for example, by power MOSFET's, to make use of their superior characteristics as almost perfect electronic high-current switches. This naturally increases the saturation resistance and the IP3, but we now need quite a few Volts for the turn-on voltage amplitude to switch these components on and off rapidly enough. Unfortunately, this circuit also has the passive mixer's attenuation at least 6dB.

b) In contrast, if we do not wish to dispense with mixing amplification, because we are actually using active

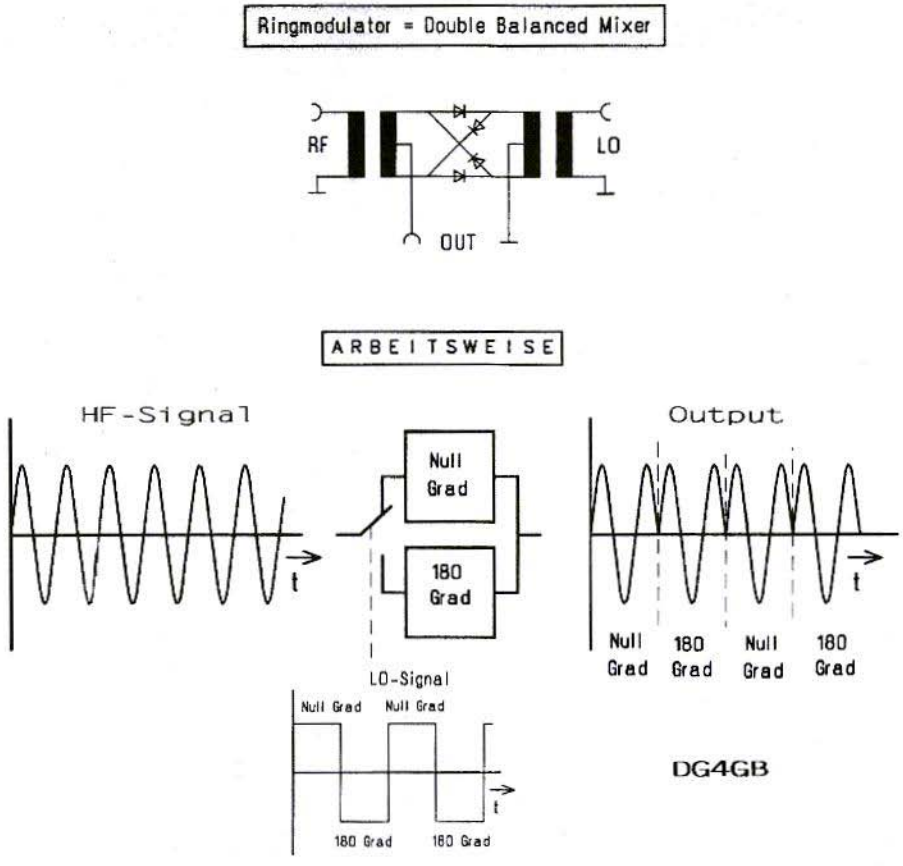
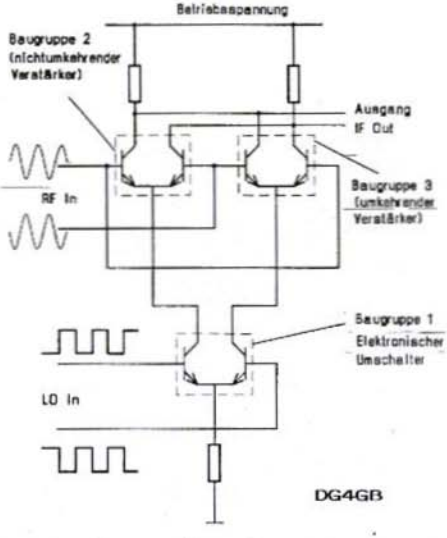


Fig.69: Principle and Functioning of a Ring Modulator
***Arbeitsweise* = Functioning, *Null* = Zero, *Grad* = Degrees**

components, then we will need three differential amplifiers in all. The way they interact can easily be seen from Fig.70. For assembly 1, we actually need only switch the power supply voltage over to fit in with the rhythm of the LO signal, thus putting either amplifier 2 (without phase reversal) or amplifier 3 (with 180° phase rotation) into operation alternatively. These two amplifier stages are therefore connected up in parallel at the input and output for the high-frequency signal path.

In the English-language literature, this layout is called a Gilbert cell. It has recently become an established constituent part of, for example, all radio phones and mobile phones, in a frequency range going far beyond 2 GHz.

The only disadvantage it has proved to have so far is the maximum level control, characterised by the IP3. Even for the best designs (with suitably high levels of zero signal current), at app. 32dBm this is still below the best brands, which are set, through the



N.B.: All assemblies for setting operating point have been omitted!

Fig.70:
Principle of the Gilbert Cell
***Baugruppe 2 (nichtumkehrender Verstärker)* = Assembly 2 (non-reversing amplifier),**
***Baugruppe 3 (umkehrender Verstärker)* = Assembly 3 (reversing amplifier),**
***Baugruppe 1* = Assembly 1,**
***Elektronischer Umschalter* = Electronic change-over switch,**
***Betriebsspannung* = Operational voltage, *Ausgang* = Output**

pedigree passive ring modulators referred to, with diodes or power MOSFET switches (giving IP3 = +40 to 45dBm).

To sum up, we can say that the multiplicative mixer has made the breakthrough in terms of professional applications where the spectrum purity of the output signal has to be high.

The industry has recently become able to supply a very wide range of suitable circuits and finished modules, the SMD versions in particular being more advantageously priced (for the same quality). There are also numerous data manuals and application manuals, with excellent articles explaining the principles involved and notes on applications [1], [2] and [3].

10.4. Examination of Diode Ring Mixer Used

Here, once again, we begin with a short list of the advantages of this type of mixer:

a) High level control, characterised by a high IP3 value.

b) Universal module, since it is suitable for use as a mixer, modulator or de-modulator for AM or SSB, a phase detector or a controllable attenuator.

c) No additional current supply required, since this is a passive component.

d) Enormously wide bands can be covered for all inputs (high-frequency and LO inputs run through small magnetic core transmitters, which means there is a lower limiting frequency of app. 20 MHz there, whilst the upper limiting frequency of this transmitter is app. 2 GHz).

e) High-frequency and intermediate-frequency outputs may be transposed, which now leaves only high-frequency signals from DC up to 2 GHz to be high-mixed.

f) Automatic matching in 50Ω systems.

But there are disadvantages as well:

a) As a passive component, the ring modulator displays at least 6 to 7dB attenuation with diodes.

b) This not only means we need additional amplification in the system, it means the noise factor in this assembly also increases by this value.

c) The turn-on voltage required (LO signal) must have a very high minimum amplitude. Depending on the type of mixer, this means an LO level of between + 7 and + 25dBm.

d) The matching, expressed, for example, in terms of the VSWR ratio, is not especially good, for in the connecting stretch between the high-frequency input and the intermediate-frequency output the transmitter must always be taken into account, together with the dynamic resistances of the diodes. The diodes are switched, and so the resistances in the signal path are continually changing and change again if the LO level fed in is increased or reduced!

e) Meticulous attention must be paid to obtaining a broad-band 50Ω adapted load, as near perfectly as possible, at the intermediate-frequency output, since otherwise for example, if a quartz filter is directly connected all signals located outside the filter's transmission band are reflected back into the mixer. They "wander around" there like ghosts, with unknown phase positions. In certain circumstances, they re-mix, and thus almost always worsen the hard-won high attenuation values for the undesirable signal fractions!

f) In professional systems, connecting up broad-band attenuators (usually 3dB)

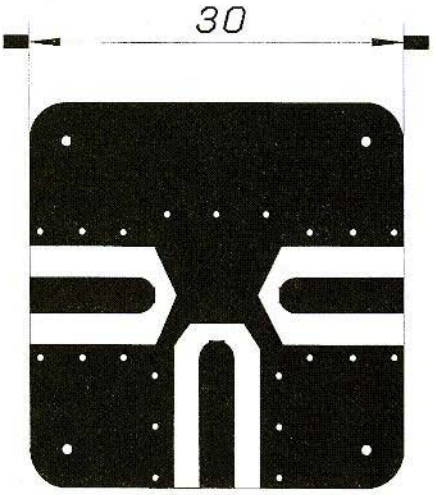


Fig.71: Board Layout for Mixer Test Structure

The recommended application range extends beyond 2 GHz; + 7dBm should be allowed for as the LO level.

Inside this cast resin cube are two ring cores for the necessary transmitters and a quartet of Schottky diodes. These components are mounted on a little printed circuit board, the three connection points of which are connected with the exterior. 50Ω microstrip lines are now connected up to these on the board.

With the layouts used so far, this has not proved a problem. The strip line width is 2.92mm.for FR4 material with a thickness of 1.5mm at a frequency of 1.7 GHz. Naturally, the required feedthroughs from the continuous earth surface to the individual "earth islands" of the layout should once again be created using silver-plated compression rivets (d = 0.8 mm.). The finished board is fastened into a milled aluminium

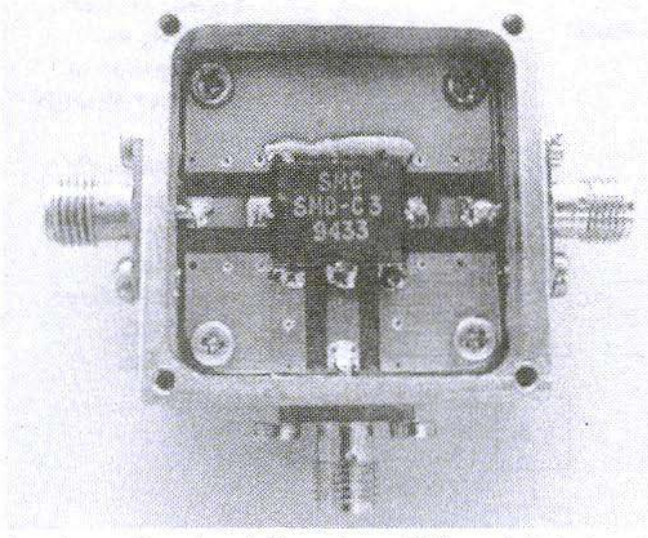


Fig.72: Photograph of Prototype

housing, with external dimensions of 35 x 35 x 15 mm, with four nickel-plated M 2.5 x 6 cylindrical screws.

The SMA sockets are precisely positioned in such a way that their neutral conductors lie just on the 50Ω microstrip lines, and can be soldered on with a little solder containing silver, without any problem. In order to avoid irregularities, all neutral conductors are chamfered below 45° with a fine file before being soldered on.

Nor should we forget to drill a "matching" hole for the neutral conductor through the housing wall. With a conductor diameter of 1.27 mm., and for 50Ω, this gives a hole diameter of 2.9mm.

Fig.71 shows the board layout. Fig.72 gives us a look at the prototype, ready to operate, but with the lid off.

Now to the electrical properties. The following characteristics are of interest to users. Most serious manufacturers include them in the data sheet:

a) Extent and frequency range of conversion loss or amplification

b) 1dB compression point

c) IP3

d) SSB noise factor

e) Insulation of various ports from one another, i.e. a dimension for the reciprocal crosstalk

f) VSWR or reflection factors at individual ports

Naturally, a very extensive and expensive array of gauging equipment is required to capture all these mixer data, and not everyone has this available.

However, tests have been carried out on mixers, the most important being the spectrum analyser with a small signal frequency response. The gauging rig shown in Fig.73 was somewhat modified as required.

The two standard signal generators represent the high-frequency and LO signal sources. The output level must be

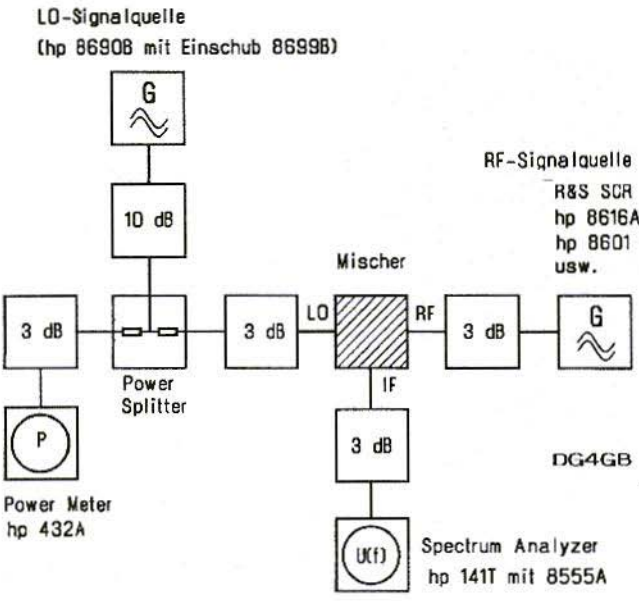


Fig.73: Measurement Rig for Investigation of Mixer Module

Signalquelle **= Signal source,**
Mit Einschub **= With insert,**
Mischer **= Mixer,**
Mit **= With,**
Usw. **= Etc.**

stabilised, and it must also be possible to alter it within precise limits. The attenuators at all the mixer ports ensure correct broad-band tuning.

10.4.1. Conversion Loss

As is well-known, when the high-frequency signal is converted in a mixer, the energy is divided into a lower and an upper sideband component frequency, namely into the total frequency and the difference frequency. Since both these fractions are of equal size, each contains half of the high-frequency energy. Thus with an ideal mixer the conversion loss is 6dB.

In practice, because of unavoidable losses in transmitters and diodes, the value is higher. Moreover, there is another big problem in determining data sheet specifications, as the mixer manufacturers are apparently not fully in agreement among themselves:

a) Some work at a constant intermediate frequency (in point-to-point radio systems, for example, this is 70 or 140 MHz). The high frequency and the LO frequency are increased until the amplitude of the intermediate frequency output signal falls by 3dB, or the conversion loss increases by 3dB. The frequency interval of the two signal generators must naturally correspond to the intermediate frequency and is kept constant.

This represents a feasible, practically-orientated method of designing a receiver with a low intermediate frequency or a reception converter, and it corresponds to the operating case of the downward mixer.

b) In the second option, the high frequency is kept constant (for example,

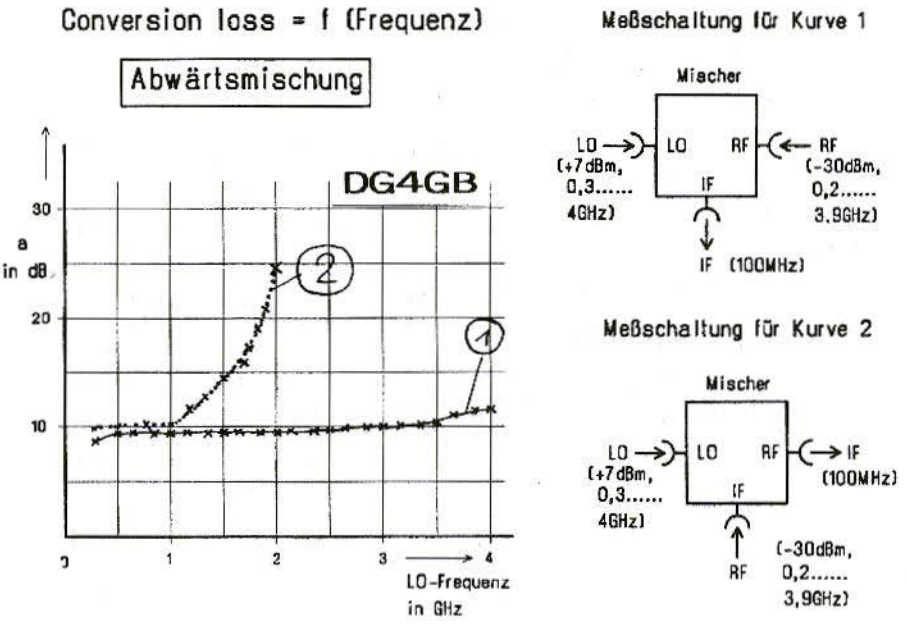


Fig.74: Downwards Mixing; *Abwärtsmischung* = Downwards mixing, *Frequenz* = Frequency, *Meßschaltung für Kurve* = Measurement circuit for curve, *Mischer* = Mixer

HF = 100 MHz), and the LO frequency is increased until the amplitude of the total or difference frequency arising decreases by 3dB or the conversion loss increases again by 3dB. This corresponds to operation as an upward mixer, as is done in SSB transmitter frequency synthesising, or in receivers with a high first intermediate frequency. In order to be clear on the behaviour of this circuit, both options are systematically examined below.

A) Operation as downward mixer:

Here a constant intermediate frequency of 100 MHz was selected. So there must always be a frequency spacing of 100 MHz between LO and high-frequency signals, and they are increased together until the conversion loss becomes unacceptably great. The high frequency is lower than the LO frequency here.

The precise operational values here, based on the mixer inputs, are:

LO signal: 300 MHz to 4 GHz, with a level of +7dBm at the LO port

HF signal: 200 MHz to 3.9 GHz, with a level of - 30dBm at the HF port

The intermediate frequency observed was 100 MHz. Fig.74 shows the conversion loss frequency response. The curve is so flat that this low-cost component can actually be used at up to 4 GHz!

As part of the trial, the intermediate-frequency and high-frequency ports were

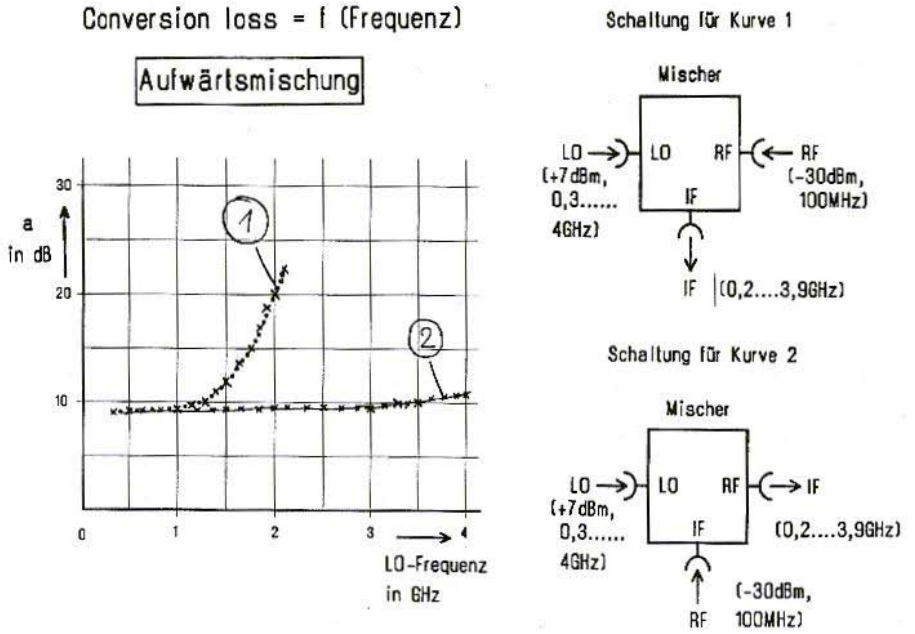


Fig.75: Upwards Mixing

***Aufwärtsmischung* = Upwards mixing, *Frequenz* = Frequency, *Schaltung für Kurve* = Circuit for curve, *Mischer* = Mixer**

also swapped round once, and the measurement was repeated (see also Fig.74). The result is quite shattering, for now the conversion loss is already 10dB greater at 1.5 GHz, and is more than 20dB greater at less than 2 GHz.

B) Operation as upward mixer:

Now a constant high frequency of 100 MHz is fed into the mixer, and the LO frequency starts at 300 MHz and is increased in 100 MHz steps.

The following operating conditions were maintained:

LO signal: 300 MHz to 4 GHz, with a level of +7dBm at the LO port

HF signal: 100 MHz, with a level of -30dBm at the HF port

The lower sideband component frequency (thus LO-HF) observed went from 200 MHz to 3.9 GHz.

As our curiosity had been aroused by the result of the first trial, we swapped the high-frequency and intermediate-frequency ports round here as well after the first pass, and then we repeated the measurements. The results can be seen in Fig.75.

There was a big surprise here too.

A circuit which can be used up to 4 GHz, with constant and relatively small conversion loss, is obtained only if the "high-mixing" HF signal of 100 MHz is fed into the intermediate-frequency port. However, if the above downwards mixer circuit is used by mistake, the mixer can be used at up to just 1.5 GHz. If the

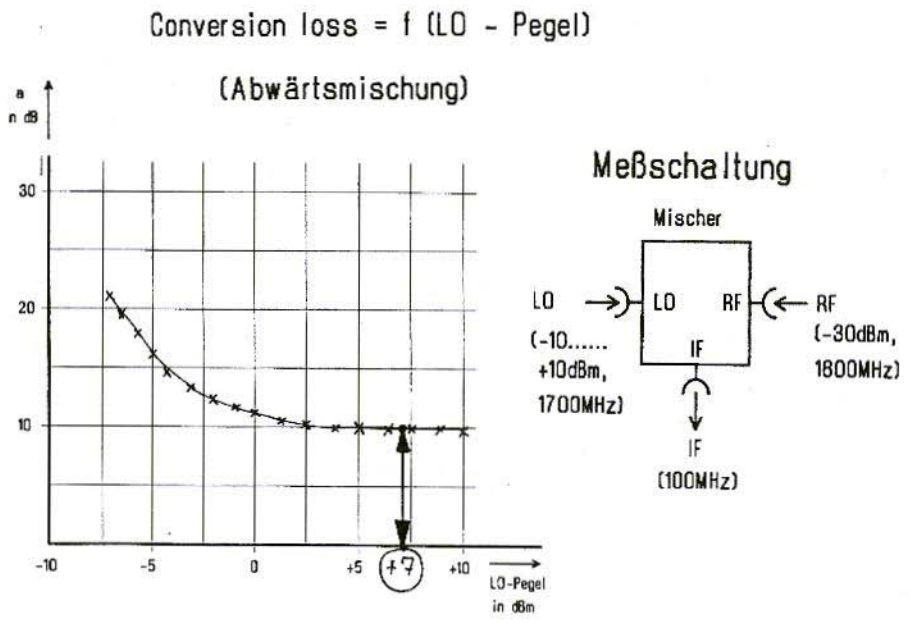


Fig.76: Relationship Measured between Conversion Loss and LO Level with Constant HF Signal; *Aufwärtsmischung* = Upwards mixing, *Meßschaltung* = Measurement circuit, *Mischer* = Mixer

frequency is increased any further, the conversion loss immediately takes on astronomical proportions.

10.4.2. Relationship Between Conversion Loss and LO Level

It was something that was being hinted at back in the chapter on principles. The faster and the more abruptly the diodes are switched through the LO signal, the better the mixer will be. In this context, this means better with respect to the conversion loss and the undesirable mixing products generated.

Fig.76 now demonstrates the relationship measured between the conversion loss and the LO level for a constant HF level.

The curve was obtained under the following conditions:

LO signal: 1,700 MHz, with a level of -10 to +10dBm at the LO port

HF signal: 1,800 MHz, with a level of -30dBm at the HF port

The difference frequency observed was 100 MHz. It can easily be recognised from this that the manufacturer's note "Please select the LO level at + 7dBm" is certainly absolutely right. If the LO level is increased any higher, this brings no further improvements, at least with this HF amplitude and with respect to the conversion loss.

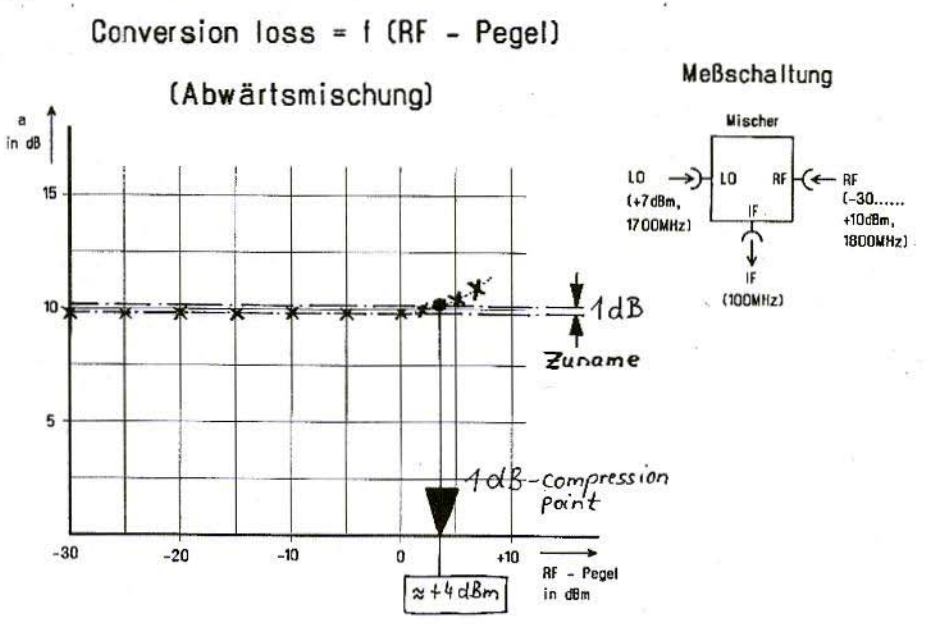


Fig.77: Measurement Curve with 1dB Compression Point: Curve shows Conversion Loss plotted against HF Level
***Abwärtsmischung* = Downwards mixing, *Meßschaltung* = Measurement circuit, *Mischer* = Mixer, *Pegel* = Level**

10.4.3. The 1dB Compression Point

If we assume that the LO level of +7dBm has been correctly selected, we can see the consequences of changing the HF level.

Fig.77 shows the measurement results for the following operating case:

LO signal: 1,700 MHz, with a level of +7dBm at the LO port

HF signal: 1,800 MHz, with a level of -30 to + 10dBm at the HF port

The difference frequency of 100 MHz was observed. The marked point at which the conversion loss increased by 1dB can (I hope) be clearly seen. This "1dB compression point" is important for three reasons:

a) This is where the saturation or limitation begins. From this point onwards, there is no point in increasing the HF level any further, because the intermediate-frequency output level no longer alters in line with the change, but very quickly becomes almost constant again.

b) The inter-modulation signals generated rise rapidly from this point onwards, and so clearly visibly that the spectrum analyser connected at the intermediate-frequency output suddenly begins to show a "lattice" of interference lines.

c) However, if we look more closely at the design manuals already mentioned, in [5], for example, we can find this:

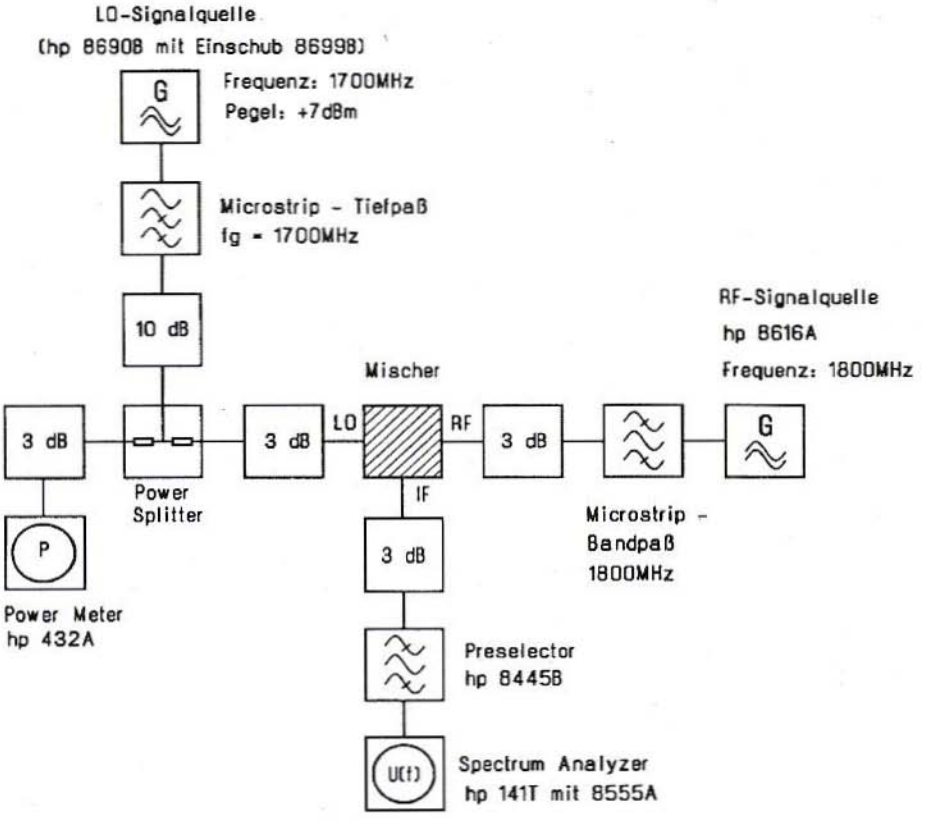


Fig.78: Extended Measurement Circuit for Determining IP
***Signalquelle* = Signal source, *Mit Einschub* = With insert,**
***Frequenz* = Frequency, *Pegel* = Level, *Tiefpaß* = Low pass,**
***Mischer* = Mixer, *Bandpaß* = Band pass**

"If about 15dB is added at the 1dB compression point at the lower band end, this gives the approximate value of the third order intercept point, IP3. The value to be added to the upper band end should then be reduced to 10dB."

10.4.4. The Intercept Point

The biggest problem here is probably the measurement rig needed to determine the values.

The intercept point itself is actually a fine thing. It supplies very accurate information concerning the distance between the undesirable interference signals arising and the desirable intermediate-frequency signal at a specific HF level. The formulation often heard of in connection with receivers, "inter-modulation free dynamic range" is relevant here as well.

Here is a short summary of the theoretical and physical basic principles:

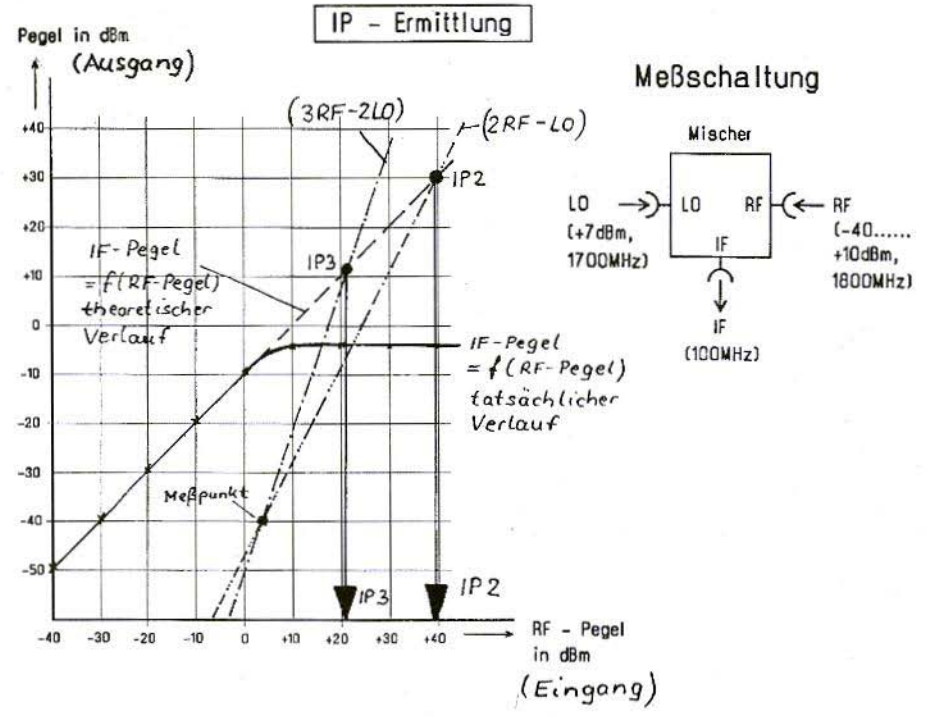


Fig.79: Graphic Determination of IP's
***Pegel* = Level, *Ausgang* = Output, *Theoretischer Verlauf* = Theoretical gradient, *Tatsächlicher Verlauf* = Actual gradient, *Meßpunkt* = Measurement point, *Meßschaltung* = Measurement circuit, *Mischer* = Mixer, *Eingang* = Input**

As soon as the HF level fed into the mixer is increased, the desirable intermediate-frequency output level also increases, up to saturation. Unfortunately, harmonic mixing on non-linear characteristics leads to many new combination frequencies, namely the notorious intermodulation signals. And this is where things get really unpleasant. Although their amplitude is very small at first, they increase much faster than the intermediate-frequency level when the HF input level rises. At some point (if no saturation takes place), this undesirable fraction becomes as large as the amplitude of the desirable total frequency or difference frequency, and this level intersection point is the famous IP!

It should also be pointed out that the rise becomes steeper and steeper as the higher-order harmonics are mixed in.

The following conditions apply:

If the HF level is increased by 10dB, then:

a) Second-order signals increase by 20dB. The annoying frequency here is usually (2 x HF - LO).

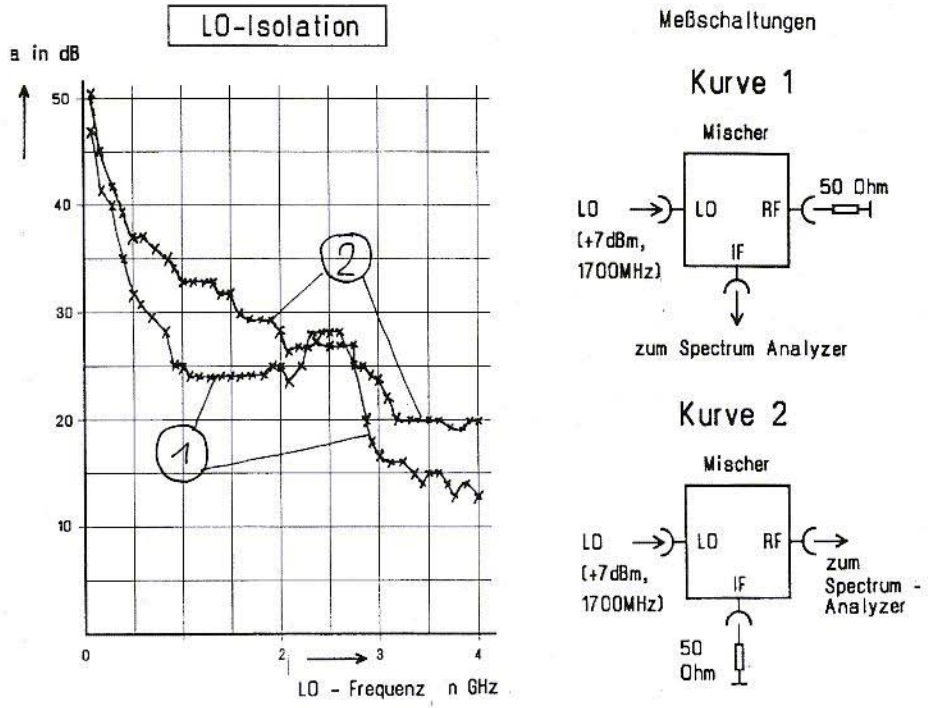


Fig.80: Measurement of Isolation of LO Signal
***Frequenz* = Frequency, *Meßschaltungen* = Measurement circuits, *Kurve* = Curve, *Mischer* = Mixer, *Zum* = To**

b) Third-order signals increase their amplitude in this case, but by 30dB. The frequency (3 x HF - 2 x LO) is particularly unpleasant here.

Signals of still higher orders normally cause no interference, since their amplitudes usually remain insignificantly low, as long as we keep far enough away from mixer saturation.

But be careful. The individual level intersection points, IP2 and IP3, are not identical!

There are particular problems here with IP3, for the third-order characteristic rise is unfortunately three times as great as for the intermediate-frequency signal. As a result, the associated inter-modulation products increase to a correspondingly great extent.

This is particularly unfavourable in practice if the mixer for example, in a receiver has too high an input level, and the HF signal consists of a fully occupied amateur band!

In addition, of all places for combination frequencies of this kind to be, they are often just at the place where the desirable intermediate frequency also emerges. For can a human being or a filter keep wanted signals and interference signals apart?

So now to the measurements.

Mixer manufacturers operate with a "two-tone HF signal" here, consisting of two closely adjacent measurement frequencies. This group of two is now converted with the LO signal, and the result is examined on the spectrum analyser. Again, the total signal and the difference signal naturally consist of these groups of two. But in addition, we discover, in all their glory, the harmonics produced in the mixer and also converted. Their attenuation can be easily read off against the desirable intermediate frequencies. The overall name for this is "dual tone third-order inter-modulation testing", and it is a feature of the specifications of commercially available mixers.

Of course, this method of measurement is very costly, since every one of the three separate signal sources must also be freed from inherent distortion. For this is the only way to determine the fractions which arise from the mixing process. The measurement rig used here is still very extensive, and is based on the following observations (Fig.78):

a) The HF is selected at 1,800 MHz, and the LO signal at 1,700 MHz.

b) The LO level is set precisely to + 7dBm at the LO port and is kept constant.

c) Both signals are freed from inherent distortion through the filters discussed in the earlier chapters (microstrip band-pass and microstrip low-pass).

d) The signal (2 x HF - LO) is observed to determine the IP2. It is located at (2 x 1,800 MHz - 1,700 MHz) = 1,900 MHz.

e) The IP3 is determined by observing the combination frequency (3 x HF - 2 x LO). It has the value (3 x 1,800 MHz - 2 x 1,700 MHz) = 2,000 MHz.

f) The HF level is now raised higher and higher until both these new frequencies can clearly be recognised on the screen and emerge from the internal noise level of the analyser. The HF level was then advanced until the amplitudes of the two inter-modulation products were exactly the same. It only remained to set the three levels at the HF level arriving at the HF port, together with the two levels (2 x HF - LO), or (3 x HF - 2 x LO), as accurately as possible.

The results showed that with an HF level of +4dBm at the HF input of the mixer, the two signals (2 x HF - LO) and (3 x HF - 2 x LO) are exactly -40dBm.

Now to the diagram, into which we first plot the relationship determined in the previous trial between the HF input level and the intermediate frequency amplitude arising. Because of the saturation, the curve naturally goes over into a horizontal line at the end. But the ideal theoretical course, which forms a straight line, is extended until we arrive at the limits of the diagram.

Two straight lines are now drawn through the measurement result (-40dBm). Their gradients are two or three times as acute as that for the relationship already plotted between the HF signal and the intermediate frequency signal, and they intersect this curve at two points (Fig.79). These two intersections correspond to IP2 and IP3!

As conscientious researchers, we also had to determine the amplitude of the intermediate-frequency signal arising

and plot it in the diagram. It is immediately apparent how far into saturation we already are.

We thus know that:

IP2 is at about +39dBm. IP3, on the other hand, is at about + 21dBm.

Anyone who wants to can check the opinion repeated in the previous chapter on the relationship between the 1dB compression point and the IP3.

With the help of this diagram, it is now no longer a problem to determine the inter-modulation distance. We need only draw a vertical line through this level (on the horizontal axis), mark its intersections with the three straight lines (gradients 1, 2, 3), read off the distances of the inter-modulation points from the intermediate frequency point, and express the result indB.

Anyone who prefers to work with formulae, on the other hand, could easily put this geometrical relationship into a corresponding expression and use it to calculate the values desired. A still easier way is to follow the method also referred to in [6]:

a) Determine the value of IP3 either from the data sheet or by adding 15dB (lower band end) or 10dB (upper band end) to the 1dB compression point.

b) Then determine the difference indB between the HF input level applied and IP3.

c) Multiply this difference by the order of the harmonic (here: 3).

d) Subtract this result and the conversion loss from IP3. This gives us the level of the third order inter-modulation interference signal.

If this method is applied to the HF level used for all measurements, -30dBm, this gives us:

a) Measured value: IP3 = +21dBm

(The rule of thumb would have been somewhat less accurate: 1dB compression point of +4dBm +15dB = 19dBm!)

b) IP3 - HF = +21dBm - (-30dBm) = 21dBm + 30dBm = +51dBm

c) 51dB x 3 = 153dB

d) IP3 - 153dB - 10dB conversion loss = + 21dBm - 153dB - 10dB = -142dBm

In this operational case, the level for inter-modulation signals is thus -142dBm, which is equivalent to 112dB below the HF level. But the difference in level at the desired intermediate frequency is much more important for the user, for there the mixer's conversion loss (app. 10dB) must also be taken into account.

Thus the final value should be 102dB below the intermediate frequency signal, and this is precisely what can be seen on the spectrum analyser.

If this seems an unbelievably good result, just imagine an HF level 30dB higher. This rise unfortunately brings about an increase of 90dB in the inter-modulation products (!).

If the intermediate frequency is now -10dBm, because of the app. 10dB conversion loss, the unwanted IM3 signal unfortunately bounces up to -14dBm +90dB = -52dBm. And so suddenly the distance between the intermediate frequency signal and the interference product is only 42dB. That could just cause some difficulties.

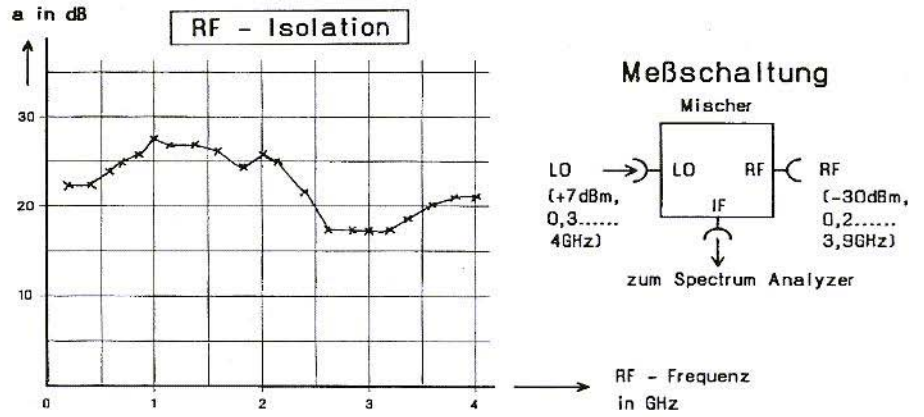


Fig.81: Measurement of Isolation of HF Signal; *Meßschaltung* = Measurement circuit, *Mischer* = Mixer, *Zum* = To, *Frequenz* = Frequency

10.4.5. SSB Noise Factor

Luckily, this is something which can be settled very quickly. As already mentioned, the intermediate frequency amplitude is reduced by the conversion loss with respect to the HF signal fed in. Thus the noise factor must rise by this amount. According to the mixer manufacturers, the SSB noise factor is calculated by adding a mere 0.5dB to the conversion loss. This gives us the noise factor we need.

10.4.6. Isolation of Ports

Nothing is perfect, and so here too we find remains of the signals fed into the mixer arriving at the intermediate frequency output port or at the opposite signal port. The difference in level between the signal fed in and the measurable residue at another port is known as isolation. If this value deteriorates too much, we can get some nasty surprises. For example, imagine a direct conversion receiver with an LO signal which suddenly starts to wander around on the antenna. The isolation behaviour should therefore also be entered in the mixer data sheet.

Fig.80 shows the isolation values of the LO signal for the frequency range 0.2 to 4 GHz, with a level of + 7dBm at the LO mixer input. At the intermediate frequency output the spectrum analyser is hanging, with the help of which we can look for the LO residues. The HF port, which is not required, is blanked off with 50Ω.

The measurement was also repeated with the HF and intermediate frequency ports swapped round. The striking features in the result are the very high conversion losses at low frequencies and their deterioration when the frequency increases. This is very unpleasant for many applications.

Fig.81 shows how high the HF signal at the intermediate frequency port can still be found to be with "correct downwards mixer operation".

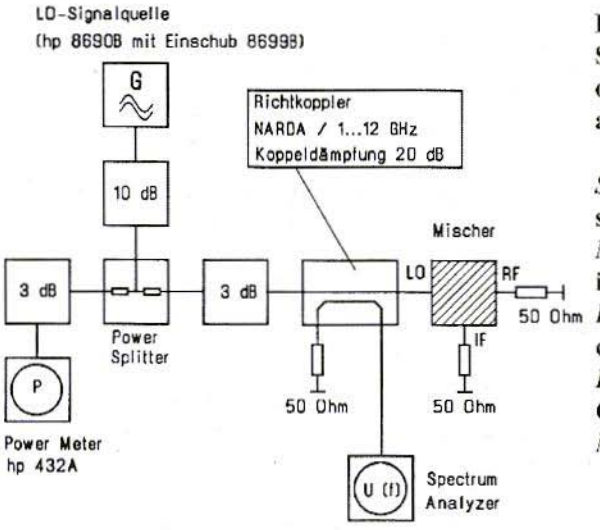


Fig.82: Experimental Set-Up for Determination of Reflection Attenuation at LO Port

Signalquelle **= Signal source,**
Mit Einschub **= With insert,**
Richtkoppler **= Directional coupler,**
Koppeldämpfung **= Coupled attenuation,**
Mischer **= Mixer**

Operational values:

LO signal: 0.3 to 4 GHz, with a level of + 7dBm at the LO port

HF signal: 0.2 to 3.9 GHz, with a level of - 30dBm at the HF port

Intermediate frequency: 100 MHz

As against Fig.80 (LO isolation), these readings show markedly fewer fluctuations and are less dependent on the frequency. But unfortunately, the downside is that they are worse results.

But be careful: the conversion loss for the intermediate frequency signal also has to be taken into account if we are trying to determine the distance between

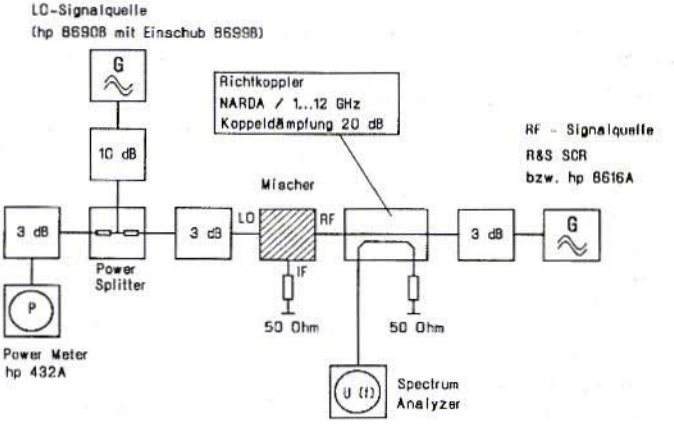


Fig.83: Determination of Reflection Attenuation at HF Port
Signalquelle **= Signal source,** ***Mit Einschub*** **= With insert,** ***Richtkoppler*** **= Directional coupler,** ***Koppeldämpfung*** **= Coupled attenuation,** ***Mischer*** **= Mixer,** ***Bzw.*** **= And / or**

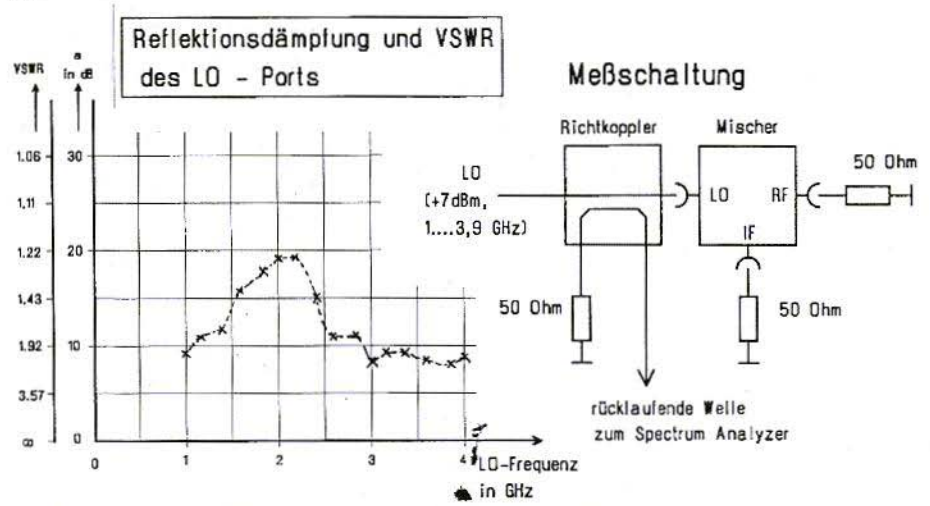


Fig.84: Measured Behaviour of Reflection Attenuation at LO Port *Reflektionsdämpfung* und VSWR des LO-Ports = Reflection attenuation and VSWR of LO port, *Frequenz* = Frequency, *Meßschaltung* = Measurement circuit, *Richtkoppler* = Directional coupler, *Mischer* = Mixer, *Rücklaufende Welle zum* = Return wave to

the desirable intermediate frequency signal and the HF residue at the intermediate frequency output, for this makes the level differences approximately 10dB worse. This is important for the layout of the subsequent intermediate frequency band pass, with reference to the inhibition attenuation required to suppress undesirable signals!

10.4.7. VSWR or Reflection Attenuation Levels for Individual Ports

Some manufacturers give only meagre information about this, or none at all, but others give precise details. However, when we look at these more closely, they provide very little encouragement, for the VSWR reaches, and sometimes exceeds, a value of 4. And no wonder, for not only the transmitter in the mixer and the diode grades, but also the LO level at which the diodes are switched, have a role to play here, as has the adapted load error at the other ports. It is therefore inevitable that attenuators or supplementary circuits should be mounted around this mixer. The mixer used here is itself not perfect, which means it imposes even higher requirements on its environment.

The measurement is somewhat costly, but the manufacturers can use the most modern network analysers for it. With these appliances, the level fed into the test object can be precisely adjusted and monitored.

This is exactly what is important, for example, at the LO port, for the correct switching through of the diodes in the mixer. It does not become really tricky unless measurements have to be made at

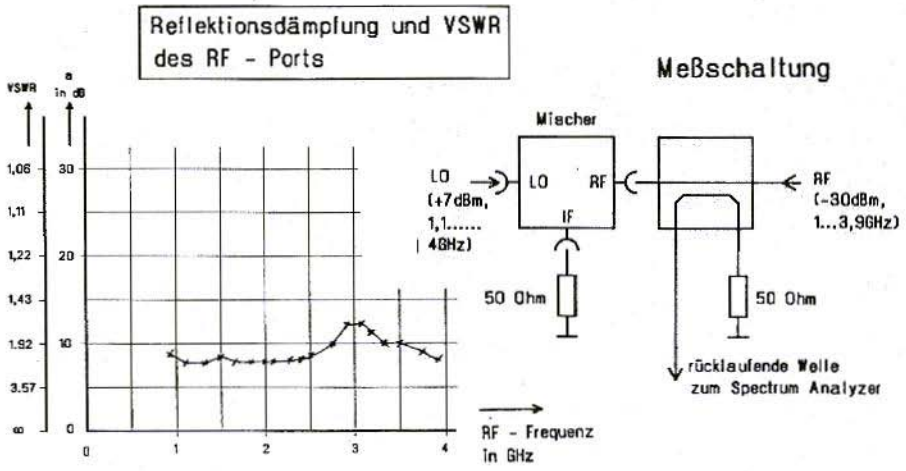


Fig.85: Measured Behaviour of Reflection Attenuation at HF Port *Reflektionsdämpfung* und VSWR des RF-Ports = Reflection attenuation and VSWR of RF port, *Frequenz* = Frequency, *Meßschaltung* = Measurement circuit, *Mischer* = Mixer, *Rücklaufende Welle zum* = Return wave to

the HF port, since then the LO port must also be selected, and both frequencies must be moved in synchronisation with one another for downwards mixer operation. This can not be done on older appliances, so we have to fall back on older methods using directional couplers (see Figs. 82 and 83). Fig.84 therefore shows this type of measurement only for the reflection attenuation behaviour at the LO port for the frequency range between 1 GHz and 4 GHz.

Fig.85 provides the reflection values required for the HF port in "downwards mixer operation", measured in the frequency range between 1 GHz and 4 GHz, with an intermediate frequency of 100 MHz. As conjectured, the results are still just usable.

10.5. Development Trends

It has already been pointed out that the integrated, multiplicative GaAs mixers are making great advances, and have driven all older designs, including diode ring modulators, blow-through mixers, etc., out of the market (and not just for mobile telephones).

The SAT receivers behind the Astra mirrors also contain only complete single-chip receivers, including PLL synthesisers for frequency ranges going up to 3 GHz. The LNC's for the frequency range from 12 GHz now contain almost exclusively the cascade stages referred to, acting as combined mixers and oscillators.

So this article will undoubtedly not be the last word on the subject of microwave mixers.

Moreover, the newest specimen IC's are already waiting on my workbench (manufacturer's specifications give the frequency range as 0.1 to 8 GHz), to be inserted into a suitable "Puff design" board. I shall be reporting on this right here at the appropriate time.

11. Literature

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[2] Communication Components / GaAs & Silicon Products
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[3] RF- and Microwave Semiconductors (Discrete, Monolithic and Multifunction) Handbook from M/A-COM

[4] Berberich, Eugen, DL8ZX
Diplexers for Ring Mixers
VHF Communications 2/96, pp. 75-81

[5] RF/IF Designers Handbook
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P. 1/6

[6] RF/IF Designers Handbook
Mini-Circuits
Chapter on "Two-tone, third-order distortion effects"
P. 1/9

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Gunthard Kraus, DG8GB

Design and realisation of microwave circuits, part 10

Development of a 2GHz oscillator

This article was originally published in UKW Berichte issue 2/1999 but it was not translated and published in VHF Communications. It has been included here because it contains details of designing inter-digital capacitors and was referenced in the article by Gunthard Kraus about a 100MHz bandpass filter published in issue 4/2003 of VHF Communications. Several subscribers requested that it was published - Andy.

1.0 Design description

1.1 The BFR92 transistor

The modern transistor BFR92 is still being produced by Siemens They provide a large number of S parameter files for their semiconductors on the Internet. It was no problem to download the appropriate "S2P" file for the operating point "4 V / 10mA", since they were arranged by collector current (rising in steps of 1 mA) and collector voltage (rising in steps of 1 V). The file used for the article below are shown in Table 1. (Note: the values for "zero Hertz" were estimated and added by hand, since PUFF would otherwise report an error if the simulation began with zero.)

1.2 Consideration of footprint capacitances

In spite of the space-saving SMD housing form, each component has two small capacitances from both "connections" to earth. These capacitances have more effect on an oscillator than on an amplifier circuit. It is therefore initially necessary to get a general idea of the order of magnitude of these additional capacitances. This can be calculated with a pocket calculator using the familiar formulae for a plate capacitor.

1.3 Important circuit details

The zero signal current of the transistor should be 10mA. It is usually set or stabilised with a basic voltage divider and an emitter resistance. Here we must take into account that the voltage divider is selected to be as high impedance as possible, since it represents an additional attenuation.

For a selected zero signal current of 10mA, the two divider resistances have the values 3.3kΩ and 3.9kΩ. They are entered in the AC circuit diagram as a parallel circuit from the base to earth.

The C2 capacitor from the emitter to earth should have as little effect on the emitter resistance as possible, otherwise the desired effect of S11 exceeding 1 and

Table 1: S parameters for BFR92.

! F	S11		S21		S12		S22	
! GHz	MAG	ANG	MAG	ANG	MAG	ANG	MAG	ANG
0.00	0.716	0 23		180	0	90	1	0
0.010	0.7158	-5.0	22.883	175.8	0.0022	85.2	0.9840	-2.2
0.020	0.7068	-9.8	22.641	171.6	0.0048	84.0	0.9756	-4.6
0.050	0.6847	-24.1	21.791	160.6	0.0124	78.8	0.9433	-11.2
0.100	0.6008	-45.1	19.271	144.3	0.0221	71.0	0.8515	-19.5
0.150	0.5134	-62.3	16.401	131.0	0.0298	65.3	0.7581	-24.7
0.200	0.4474	-76.5	14.016	121.9	0.0356	62.9	0.6806	-27.2
0.250	0.3892	-88.9	12.039	114.4	0.0402	62.0	0.6199	-28.5
0.300	0.3444	-99.3	10.482	109.0	0.0450	62.3	0.5780	-28.8
0.400	0.2906	-115.9	8.269	101.1	0.0540	62.3	0.5229	-28.4
0.500	0.2581	-131.0	6.764	94.2	0.0630	63.7	0.4861	-28.1
0.600	0.2396	-142.6	5.739	89.2	0.0722	64.4	0.4639	-27.5
0.700	0.2239	-154.2	4.986	84.4	0.0815	65.0	0.4504	-27.5
0.800	0.2223	-163.3	4.384	80.4	0.0912	64.8	0.4405	-28.0
0.900	0.2152	-172.1	3.923	76.6	0.1011	64.8	0.4321	-28.7
1.000	0.2172	178.7	3.564	73.4	0.1102	64.6	0.4244	-29.5
1.100	0.2222	172.2	3.281	70.1	0.1204	64.4	0.4194	-30.8
1.200	0.2273	165.9	3.045	66.8	0.1301	64.1	0.4113	-31.5
1.300	0.2274	160.3	2.821	64.0	0.1404	63.3	0.4067	-32.8
1.400	0.2318	154.4	2.627	61.7	0.1494	62.8	0.4023	-34.1
1.500	0.2357	151.0	2.486	58.9	0.1594	62.0	0.3976	-35.7
1.600	0.2513	145.5	2.330	55.9	0.1695	61.2	0.3935	-37.1
1.700	0.2550	140.8	2.214	52.8	0.1792	60.5	0.3866	-38.6
1.800	0.2663	136.1	2.125	50.9	0.1890	59.5	0.3826	-40.4
1.900	0.2791	134.3	2.039	48.3	0.1987	58.6	0.3767	-42.4
2.000	0.2838	130.6	1.935	45.7	0.2089	57.6	0.3708	-44.1
2.200	0.3025	124.4	1.784	41.1	0.2289	55.7	0.3584	-48.4
2.400	0.3260	118.1	1.663	36.0	0.2487	53.6	0.3424	-52.8
2.600	0.3530	113.2	1.570	31.2	0.2686	51.4	0.3258	-57.6
2.800	0.3742	108.8	1.473	27.1	0.2883	49.1	0.3083	-62.7
3.000	0.4003	105.3	1.403	22.5	0.3079	46.7	0.2914	-68.9
3.500	0.4617	96.7	1.236	13.3	0.3539	40.4	0.2438	-88.5
4.000	0.5198	89.4	1.105	4.1	0.3994	34.2	0.2151	-115.5
4.500	0.5713	83.2	1.007	-3.8	0.4430	27.5	0.2102	-146.2
5.000	0.6123	77.4	0.916	-12.1	0.4799	20.4	0.2334	-174.0
5.500	0.6586	71.3	0.803	-19.3	0.4958	12.6	0.2400	159.3
6.000	0.6582	66.2	0.772	-22.2	0.5416	8.6	0.3341	154.1

! SIEMENS AG Semiconductor Group, Munich

thus negative input resistance is impaired. So a small SMD RF choke of 22nH is wired in series with the 82Ω resistor. A series loss resistance of 10Ω is added, in accordance with the manufacturer's specification (Q=50) Its self resonance is taken into account by a capacitance of 0.25pF.

The output from the oscillator is taken through a 3dB attenuator on the collector for broadband matching. The DC voltage drop caused by the collector resistance of 56Ω is compensated for by a parallel circuit with a 22nH choke.

1.4 AC circuit diagram for simulation with PUFF

Now let's gather together the elements still missing for the simulation.

For the ceramic resonator, the newest Internet documents from Siemens are consulted. For the selected type of resonator, with a cross section of 6 x 6 mm.^{2} and $\varepsilon = 38$, the quality is Q = 600, and is predominantly limited by the ohmic losses of the silver coating and the skin effect. We thus consider the resonator as a lossy transmission line. The following data in PUFF component list F3:

ql 12.3Ω 90° 600Qc

The resonator is coupled to the transistor circuit through an inter-digital capacitor, which impairs the operating quality and thus the noise characteristics of the oscillator if it is selected to be too big. It's value lies somewhere in the vicinity of 0.2pF.

With this solution, we must not overlook the fact that, due to the inter-digital structure, additional inherent capacitances to earth arise. They are very dependent on the board thickness and the actual dielectric constants of the board material. For this design, we simply start from a capacitance value in the order of the coupling capacitance required (0.2pF) on each side.

As regards the two capacitors, C1 and C2, of the capacitive voltage divider, the value provided of 1pF each is used for the first draft. But in the next section we shall investigate its influence and optimise the circuit.

Thus, taking all the details mentioned into account, we arrive at a lavishly detailed configuration for the PUFF simulation, Fig 1. The dotted line shown here between points A and B is an imaginary separation between the resonator section with the coupling capacity and the active part of the circuit, which produces the negative input resistance required for the permanent oscillation. We can thus now vary different component values in the transistor circuit with PUFF's "component sweep" and deter-

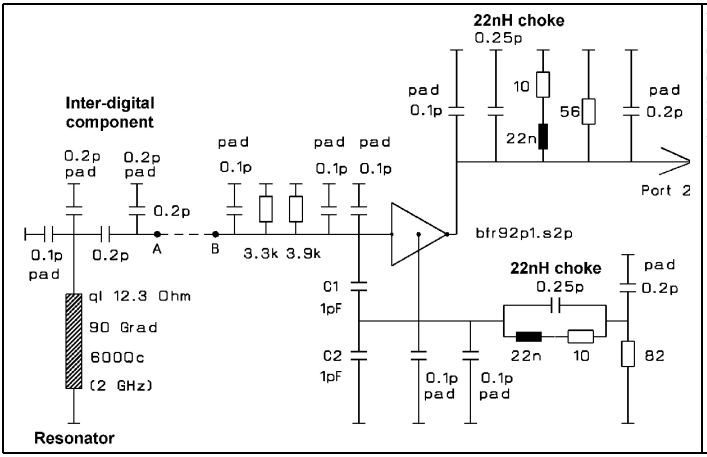


Fig 1: AC circuit diagram of oscillator for simulation with PUFF.

mine their influence on the S11 factor of the circuit (i.e. at point "B").

1.5 Definition of C1 and C2

With the help of PUFF and its "component sweep", the optimal values of the two capacitors are determined, one after another. "Optimal" means that we are trying to obtain the highest (positive) value for the input reflection, S11, at f = 2GHz. To do this, we proceed as follows:

Step 1:

First, the components required for the amplification circuit from point B alone are entered in list F3 (to the right of the separation line in Fig. 1) and the corresponding simulation circuit is created in field F1.

Step 2:

Likewise, a question mark is entered under capacitor C2 in F3 (component "e") in front of the actual component value, with 1pF. The other settings such as "design frequency 2GHz, board thickness = 1.5mm, ε = 4.3, copper thickness = metal thickness mt = 0.035 mm, tangent of loss angle = loss tangent = 0.02" have hopefully already been entered in the file. If necessary, call up the file used with a text editor and check the entries!

Step 3:

After pressing on F2, we first alter the axial division for the "swept capacitor C2" from 0 to 2pF on the horizontal axis of the lower diagram ("rectangular plot").

Step 4:

If we start the plotting process (Fig 2), then we very soon notice that we obtain a maximum value for C2 = 0.8pF at S11. We enter this value in list F3 against component "e", once we have removed the question mark again.

Now the question mark is put against the value for capacitor C1 in F3 (component "m") instead. Fig. 3 gives the result of the plotting process. One very satisfactory observation is that the value of C1 = 0.8pF (as in C2) precisely matches, and produces the maximum value of S11.

1.6 Determination of coupling capacitor required

First, we supplement the circuit in field F1 by adding the resonant section, including the inter-digital structure with the pad capacitances. But we still do not have the constant current source "cs.dev"

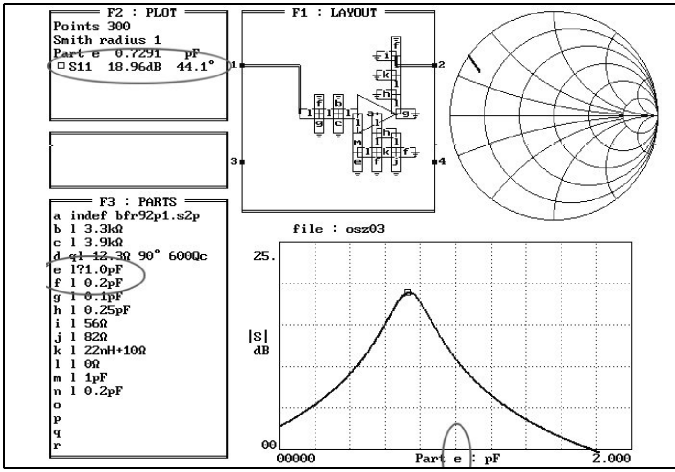


Fig 2: Determining the optimal value of C1.

already referred to. It is triggered by port 1, and feeds its output signal directly in at the resonator connection.

If we now set the radius of the Smith chart on 30 and plot the forward amplification S21 in the area between 1.8 and 2GHz (Fig. 4), we now need only look at the direction in which the S21 circuit on the Smith chart is curved.

A curve to the left means that the point of the oscillation application has already been reached, and even passed. The resonator has consequently already been attenuated again through the "excess negative resistance", and S21 has thus decreased. The only remedy against this is to make the coupling capacitor smaller.

A curve to the right means that the negative input resistance of the transistor circuit is not yet sufficient to cover the losses. The oscillator is therefore still oscillating, and the coupling capacitor must be enlarged.

At the precise application point of the oscillation, we obtain an infinite amplification, S21 – this point must be searched for with the aid of PUFF.

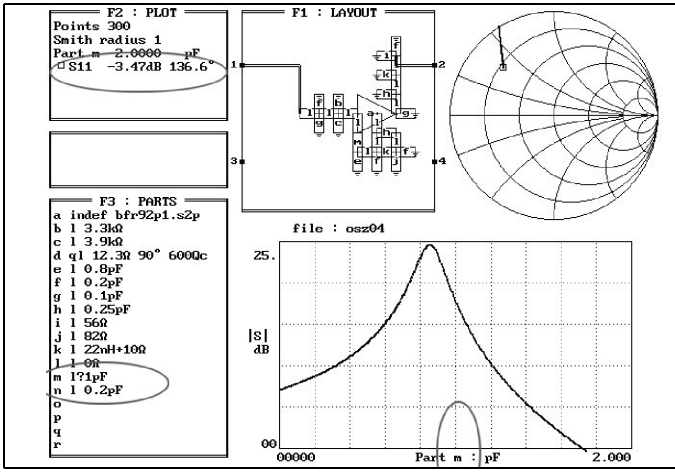


Fig 3: Determining the optimal value of C2.

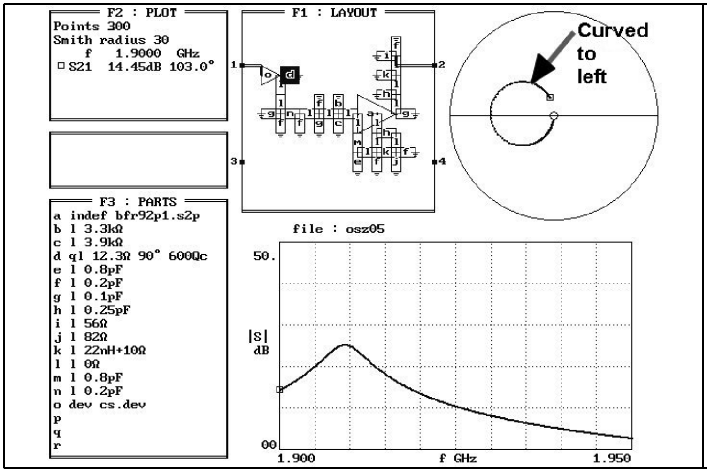


Fig 4: First simulation run for the project.

As can be seen from Fig. 4 and from the Smith chart curve to the left, the oscillator is oscillating for a long time and the coupling must be reduced. Fig. 5 shows the necessary changes to the coupling capacitor (component "n") to find precisely the application point we are searching for, with S21 > 100 dB.

Release the earth connection on the resonator, connect the resonator end which this makes accessible to port 1, and plot the input reflection S11 for the area around the resonance frequency, with a high resolution. The application point for the oscillation lies at the point where S11 is precisely zero and the phase changes from "-180 degrees" to "+180 degrees". If this "check simulation" is carried out very precisely, I'm pleased to say that we obtain precisely the same frequency value as with the new method.

As already stated, it does not make very much sense to select this operating condition. The smallest scatter or change leads to the breakdown of the oscillation or prevents oscillation from starting. So the positive feedback is set to be stronger than is necessary, and then initially, through a suitable amplitude stabilisation, we increase the oscillator voltage,

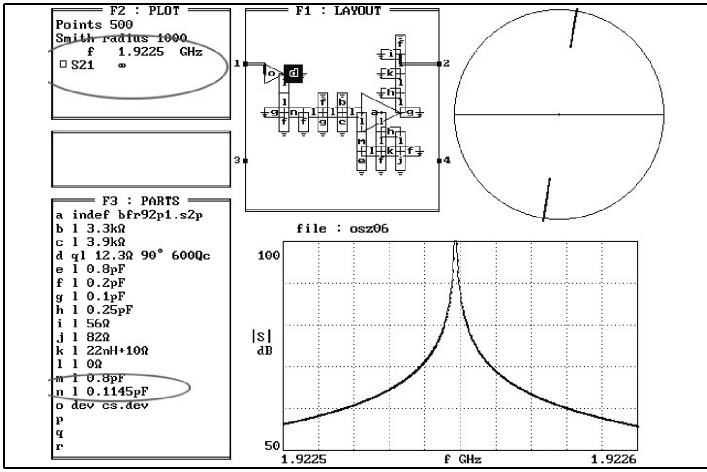


Fig 5:Determining the point of oscillation by changing the coupling capacitor at the resonator.

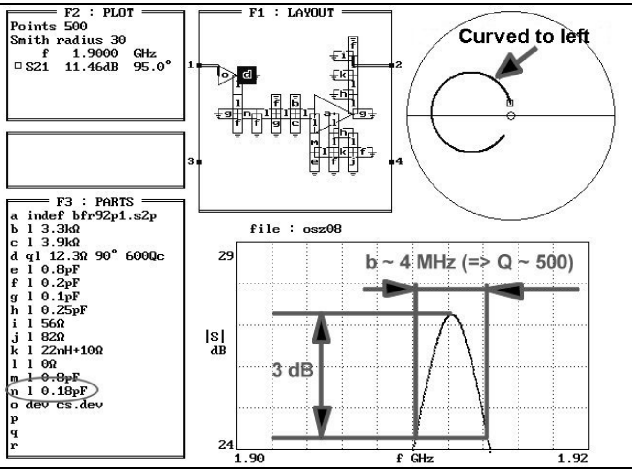


Fig 6: Definition of the coupling capacitor for a Q≥500.

and then reduce the amplification as the amplitude increases, until a stable, maximal output level is obtained under all possible operating conditions.

This can be obtained by making the coupling tighter until, through the "excess negative resistance", a circuit quality of app. Q = 500 to 1000 is set. It can then easily be recognised from the rectangular plot that the S21 curve displays a 3dB bandwidth from approximately 2 to 4MHz, in accordance with the well-known quality / bandwidth formula.

Fig. 6, finally, represents this operating case. From it, we can detect that the coupling capacitance needs to be max. 0.18pF, and should take the form of an inter-digital capacitor. As was mentioned earlier, the oscillation frequency, due to various additional capacitances in the circuit, has to be below the natural resonance of the resonator.

The simulation already leads us to suspect that we will probably essentially not exceed 1910MHz.

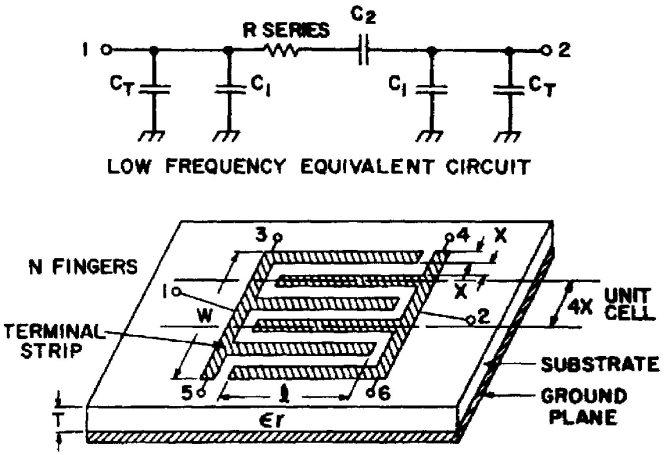
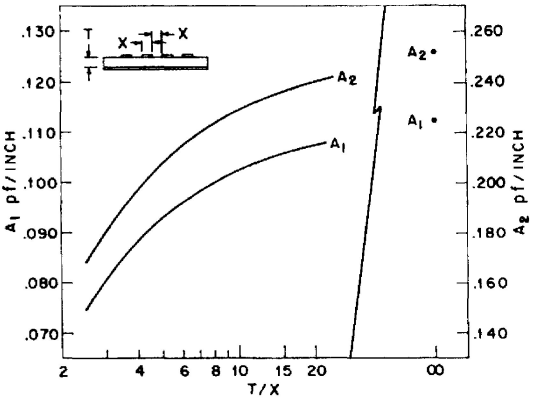


Fig 7:Geometry of the inter-digital capacitor.



Effect of substrate thickness on C_2.

Fig 8: Determining the sizes A_1 and A_2 for X=0.254mm and for a board thickness of T=1.5mm, i.e. T/X=5.9

1.7 Calculation of an inter-digital capacitor

The most important principles can be found in the article by Gary D. Alley [1]. He uses the finger structure represented in Fig. 7, in which both the finger width and the interval between fingers are described by "X" and are identical. The following relationship with the finger length then applies for the coupling capacitance (in pF):

$$C_{coupling} = \frac{L^2}{W} x(\varepsilon_r + 1)x[(N-3)xA_1 + A_2]$$

Where:
L = finger length in inches
W = width of formation in inches
N = number of fingers
ε_r = relative permittivity of board

A_1 and A_2 are two values which depend on the relationship of the board thickness (T) to the interval between the fingers referred to (X). Their values for the T/X relationship used can be taken from Fig. 8.

The inter-digital capacitor can thus be drafted for 0.18pF in the following manner.

Step 1:

As a preliminary, select, for example, a structure with 6 fingers and a finger width and/or interval between fingers, X = 0.01 inch = 0.25 mm This gives a width W for the formation of:

W=11 x 0.01 inch = 0.11 inch = 2.79mm

The relative permittivity of the FR4 material for f = 2GHz is still, according to experience, only ε = 4.3.

Step 2:

For a board thickness T = 1.5mm and an interval X = 0.254mm, there is a relationship of:

$$\frac{T}{X} = \frac{1.5mm}{0.254mm} = 5.9$$

In addition, from Fig. 8 we obtain the values: A_1 approximately 0.095 and A_2 approximately 0.22.

Step 3:

Thus the above formula can be used to calculate the values for a table for finger lengths of between 0.04 inch (approximately 1mm) and 0.12 inch

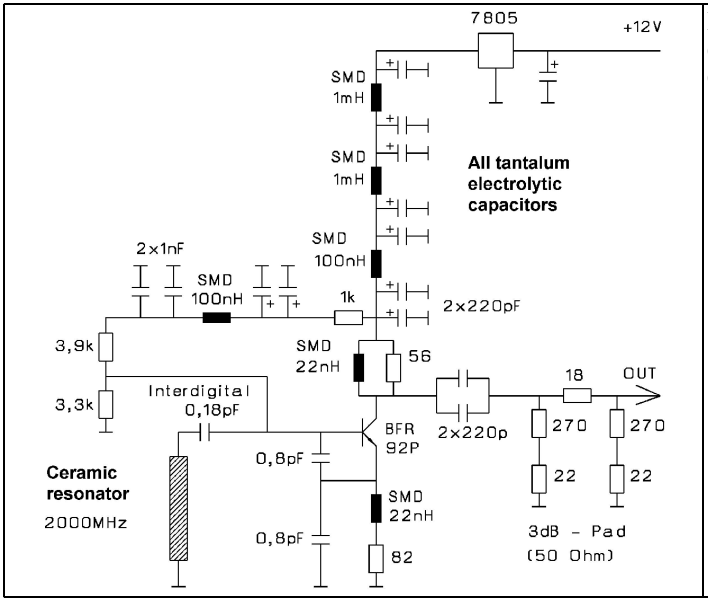


Fig 9: Circuit diagram of designed oscillator.

(approximately 3mm), with a step width of 0.002 inch (approximately 0.5mm) :

Finger length in inches	0.04	0.06	0.08	0.1	0.12
Finger length in mm.	1.016	1.524	2.032	2.54	3.048
Coupling capacitance in pF	0.0426	0.096	0.17	0.27	0.38

The rest is not a problem, for with a finger length of 0.08 inch (approximately 2mm) we have almost arrived at C = 0.18pF.

1.8 Circuit diagram and board layout

As stated in the previous section of the article, the power supply passes through a fixed voltage regulator and some expensive filters. The complete circuit is shown in Fig. 9. Fig. 10 shows the fully equipped board in the milled aluminium housing and the broadband filter for the operating voltage. The inter-digital capacitor, the ceramic resonator, the filter with the tantalum electrolytic capacitors and the attenuator at the output are easily recognisable.

1.9 Test results

The output frequency was specified at f = 1.896540GHz using a frequency counter (hp 5245). A 24 hour measurement yielded a frequency change of max. 200kHz. The frequency change is greatest in the first 5 to 10 minutes; a remedy can be provided by a temperature controlled oven.

The formation is mechanically very stable, due to the milled aluminium housing and the 3mm thick cover with “conductive silver sealing”, and so the sensitivity to feadback is very low. Mechanical shocks, e.g. from a screwdriver, can displace the frequency by a maximum of some kHz, and then it finally returns to the old value.

The use of a fixed voltage regulator for the stabilisation of the power supply

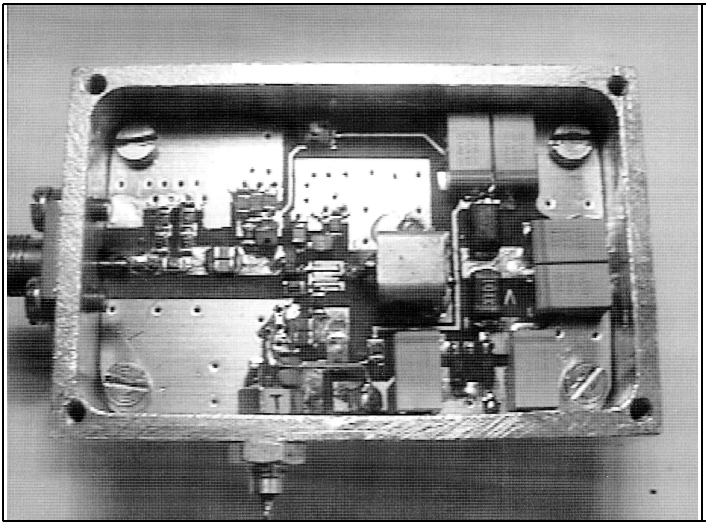


Fig 10: Completed prototype of the oscillator in a milled aluminium housing.

voltage also pays dividends If we increase the applied power supply voltage by 1V, the frequency changes only by approximately 5 kHz.

2.0

References

[1] Gary D. Alley: "Interdigital Capacitors and Their Application to Lumped-Element Microwave Circuits". IEEE Transactions on Microwave Theory and Techniques, Vol. MTT, NO. 12, December, 1970, Page 1028.

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